The Microwave Newsletter Technical Collection

Compiled by
J N Gannaway, G3YGF
S J Davies, G4KNZ

RSGB

THE MICROWAVE NEWSLETTER
TECHNICAL COLLECTION

Compiled by: J N Gannaway, G3YGF
S J Davies, G4KNZ

THE MICROWAVE NEWSLETTER TECHNICAL COLLECTION

Introduction

This collection of technical items has has been produced as a result of requests for back issues of the Microwave Newsletter. It was not until all the material had been gathered together that we realised quite how much there was, and what a wide variety of subjects had been covered.

The Microwave Newsletter was started in April 1980, as a newsletter produced just before each of the six microwave cumulative contests to circulate details of which sites people were planning to operate from. Six issues were produced, copied and distributed by the three editors in 1980. In 1981 the RSGB agreed to take over the distribution, and since then ten issues have been produced each year. Although the newsletter was originally aimed at those taking part in the 10GHz cumulative contests, it soon expanded to include items relavent to the other microwave bands.

The technical items have been extracted from past newsletters, some minor corrections made, and sorted into subjects - eg 10GHz Narrowband, Oscillator sources etc. In addition some other relevant articles have been included for completeness; for example, there have been a number of articles in the newsletter referring to the Microwave Committee UHF Oscillator board, so the original RadCom article has been included.

This collection is not intended to replace the newsletter, but rather to provide the newcomer with an easy means of getting up-to-date.

Unfortunately the cost is higher than might be expected for a publication this size. This is because demand does not warrant using anything other than photocopying, at several pence per sheet.

We hope to update this collection periodically, perhaps every 12 months, by adding new material as available.

In order to make such alterations easier, we have tried to avoid more than one subject per page, and have used what may seem a rather strange numbering system. This was to make it easy to add or remove pages or sections without disturbing the numbering of the rest of the book.

We apologise for the poor quality of print in some places. It was produced by cutting and re-pasting the original newsletter masters, as this was the easiest solution available in a reasonble time-scale. We hope to gradually replace the poorer bits when we can, by incorporating them in updates.

To subscribe to the microwave newsletter, contact the Membership Services Dept. at RSGB HQ, Alma House, Cranborne Road, Potters Bar, Herts. Mark the envelope with "Microwave Newsletter".

We would like to thank the many contributors for their material covering a wide range of interests, and all the subscribers for their support. Please keep the ideas flowing in - whether they be articles, sketches on the back of an envelope, or telephone calls!

Other microwave material appears in the monthly column, "Microwaves", (or as articles) in Radio Communication and in the microwave chapter of the RSGB VHF/UHF manual.

G3YGF and G4KNZ

THE MICROWAVE NEWSLETTER TECHNICAL COLLECTION

Covering all the technical material
contained in the Microwave Newsletter from
April 1980 to May 1983.

Compiled by Julian Gannaway G3YGF
Steve Davies G4KNZ

Newsletter Editors Julian Gannaway G3YGF
Hugh Griffiths G4CNV
Steve Davies G4KNZ

ACKNOWLEDGEMENTS

Material from the following authors has been used and we gratefully acknowledge their contributions.

G3AYJ	GM3FYB
G3HWR	G3JHM
G3JVL	G3LQC
G3NAQ	GW3PPF
G3PFR	G3RPE
G3RZD	G3UFW
G3WDG	G3WJG
G3YGF	G3YGO
G3ZIV	G4ALN
G4AUC	G4BGP
G4CNV	G4DDK
G4EBF	G4ERP
G4FRE	G4FZZ
GM4HFM	G4KGC
G4KNZ	GW4LXO
G4MBS	G6GCI
G6XM	G8AGN
G8APP	G8DEK
G8DIC	G8EXL
G8KRD	G8SHF
G8TXG	F6DLA

and also material from:

Microwave Engineers Handbook
Microwave Journal
Mullard Ltd.
Radio Communication

Contents

Introduction

Acknowledgements

1. 1.3GHz

1.1 Introduction to Equipment for 1.3GHz Stations

1.2 1.3GHz Beacons (Map and Data)

1.3 Alford Slot Aerial for 1.3GHz

1.4 1.3GHz Repeater Duplexing Filters

1.5 23cm Transverter System (Block Diagram)

2. 2.3GHz

2.1 44 Element Loop Quad Design for 2.3GHz

2.2 2.3GHz Transverter system (Block Diagram)

3. 3.4GHz

3.1 9cm High Level Mixer

3.2 Waveguide Components
3.2.1 Waveguides & Filters for 3.4GHz
3.2.2 Waveguide Transition for 3.4GHz

4. 5.7GHz

4.1 5.7GHz Transverter Designs
4.1.1 5.7GHz G3JVL Transverter
4.1.2 "Folded" G3JVL Mixer for 5.7GHz

5. 10GHz General

5.1 Introduction
5.1.1 Making a Start on 10GHz
5.1.2 An Introduction to 10GHz Operating in the Cumulatives

5.2 Relative Performance of AM/FM/SSB

5.3 Components
5.3.1 Waveguide Switch
5.3.2 Flanges

5.4 10GHz Gunn Reflection Amplifier

5.5 Aerials
5.5.1 Horn Antennas from Oilcans

5.6 Propagation
5.6.1 Use of Reflectors on 10GHz
5.6.2 Propagation Modes

5.7 Talkback

5.8 Equipment Performance on the Intermediate Bands

6. 10GHz Wideband

6.1 Gunn Oscillators
6.1.1 Varactor Tuning
6.1.2 Mechanical Tuning
6.1.3 Mounting
6.1.4 Improvements to self-oscillating Mixers
6.1.5 The Mullard CL8630 X-Band Gunn Oscillator

6.2 A Gunn Diode and Oscillator Test Rig

6.3 A Simple 10GHz Transceiver

6.4 A Doppler Tranceiver for 10GHz

6.5 Gunn Diode Power Supply

6.6 A 10.7MHz IF Preamplifier and Discriminator

7. 10GHz Narrowband

7.1 Varactor Multipliers
7.1.1 Low Power Multiplier to 10GHz
7.1.2 Improved Low Power Multiplier to 10GHz
7.1.3 Higher Powers on Narrowband
7.1.4 A Varactor Tripler from 384MHz to 1152MHz

7.2 10GHz Bandpass Filters
7.2.1 Post Coupled Filters
7.2.2 Iris Coupled Filters

7.3 The G3JVL Image-Recovery-Mixer Transverter for 10GHz
7.3.1 Original Design (waveguide components)
7.3.2 Further Information on a Complete Transverter
7.3.3 Tapered Section Mixer
7.3.4 Improvements to Multiplier and Mixer Diode Mounts
7.3.5 Alignment

7.4 Comments on the VHF Comms Narrowband Tranceiver

7.5 Frequency Markers
7.5.1 A Crystal-controlled Frequency Marker for 10GHz
7.5.2 Marker Oscillator for 10GHz

7.6 Change-over System

7.7 Beacon Building

8. 24GHz

8.1 The Plessey GDO33 Gunn Oscillator
8.1.1 A Power Supply/Modulator for the GDO33 24GHz Oscillator
8.1.2 Modifications to the GDO33
8.1.3 Notes on GDO33 PSU

8.2 24GHz Components
8.2.1 Diode Mounts
8.2.1.1 An Improved 24GHz Diode Mount
8.2.1.2 A 1N26 Mixer for 24GHz

8.3 Control of a Gunn Oscillator by Phase Locking Techniques

9. Millimetric Bands

9.1 Introduction

9.2 Propagation
9.2.1 Atmospheric Attenuation
9.2.2 Path Losses

9.3 Ideas For Getting Started

9.4 Metal sections for Use as Waveguides

10. Oscillator Sources

10.1 A High-quality UHF Local Oscillator Source
10.1.1 Constructional Details
10.1.2 Alignment
10.1.3 Stabilised Low Voltage Drop Regulator
10.1.4 Performance
10.1.5 Temperature Compensation
10.1.6 Extending the Upper Frequency Limit
10.1.7 Additional Notes
10.1.8 Oscillator Quality on 10GHz
10.1.9 Mic Preamp

10.2 Comments on VHF Comms Local Oscillators

11. Filter Design

11.1 Design Calculations for Waveguide Bandpass Filters

12. Antennas

12.1 G3JVL Flyswatter

12.2 Dishes
12.2.1 Use of Spherical Reflectors as Approximations to Parabolic Reflectors

12.3 Baluns
12.3.1 A Balun for use with 0.141" Semi-Rigid Coax

12.4 Alford Slot Antennas
12.4.1 Theory of Operation

13. Calculating Bearings

13.1 Bearing Calculations using NGRs
13.1.1 Notes on the National Grid System
13.1.2 Calculating Bearings
13.1.3 Cross Channel Bearings
13.1.4 NGR Bearing Calculation Program
13.1.5 Notes on the NGR Program

13.2 Bearing Calculations using Latitude and Longitude

13.3 Conversion Between Lat/Long and NGR
13.3.1 Conversion of Lat/Long to NGR
13.3.2 Lat/Long to NGR Conversion Program

14. Radiation Hazards

14.1 Licencing

14.2 Measurements of Power Densities

14.3 Proposed RF Radiation Safety Limits

15. Test Equipment

15.1 Power Supplies and Voltage Regulators
15.1.1 Battery Charger/PSU
15.1.2 A Low Voltage Drop, Reverse Supply Protected Regulator
15.1.3 Reverse Polarity Supply Protection

15.2 Low Power 50 Ohm Loads

15.3 Audio Voltmeter

15.4 Attenuators
15.4.1 Design and Construction of Simple Attenuators
15.4.2 Switched Attenuators

15.5 Power Meters
15.5.1 A Simple VHF/UHF Power Meter
15.5.2 A Broadband Linear RF Power Meter
15.5.3 Rough Measurements of Power
15.5.4 Power/Current Relationship in Schottky Diodes

15.6 Directional Couplers
15.6.1 A Directional Coupler using Semi-rigid Coax
15.6.2 High Power Directional Coupler for 1296MHz
15.6.3 Notes on Using Directional Couplers

16. Miscelleneous

16.1 EME Link Budget Program

16.2 Heliax Connectors using Standard N-type Plugs and Sockets

16.3 Notes on the IC202

16.4 WR/UR Waveguide Numbering system

16.5 Problems in using Droitwich-Locked Frequency Standards

16.6 UoSAT

16.7 Some pitfalls in noise figure measurements

16.8 Bibliographies
16.8.1 VHF Communications
16.8.2 DUBUS
16.8.3 Microwave Newsletter
16.8.4 Radio Communication (10GHz)

Section 1 1.3GHz

1.1 Introduction to Equipment for 1.3GHz Stations

This summary is based on G3JVL's lecture at N.E.C. To anyone new to the band it gives some general guidance and references for more information.

The basic requirements, in order of importance, are: ANTENNA, COAX FEEDER RECEIVER, TRANSMITTER, and RELAYS.

ANTENNA: Maximise the gain and provide good directional control.

This means using either a Yagi or parabolic dish. The loop yagi has a slightly higher gain than a conventional Yagi for the same length. The original 27 element loop yagi design was 8.64 wavelengths long giving a gain of 19.0dBi. The latest version has a 17 wavelength boom and gives a gain of 21.5dBi. A conventional yagi in common use is the 19 element F9FT design; an alternative design is that by DL6WU. These have a gain of around 17dBi. For information on parabolic dishes see the VHF Manual - a 10ft diameter dish would have a gain of approx 29dBi.

The beamwidth of a 24 element loop yagi is 17 degrees. The beamwidth of a 10ft dish is only 6 degrees. It is important to be able to point the antenna with an accuracy of better than the beamwidth.

The Alford Slot is another useful antenna on 1.3GHz, particularly for a beacon or repeater antenna. This gives an omni directional horizontally polarised radiation pattern with a gain 5 to 8 dBi.

COAX FEEDER: Use good quality coaxial cable.

The smallest to use is UR67 - loss 8dB/100ft. For long runs 1/2" or larger heliax is preferred. If the pre-amp is mounted at the mast head the receive feeder to the converter is not so important - UR67 would suffice for this.

RECEIVER: Use a well designed converter and low noise pre-amplifier.

There are several converters (or transverters) available commercially and several good published designs.

The pre-amplifier should be mounted at the masthead - this will improve the system performance considerably even when good coax is used. Use a GaAs FET type for lowest noise. There are some good bipolars about also - eg NE645 with NF of about 1.5dB. For tropo working aim for a noise figure of about 1.5dB. There is not much point going below about 1dB, except for EME working, because ground noise will mask the improvement.

TRANSMITTER: Aim for 10W minimum output power.

2C39's or derivatives can be used. A good design can give 50W output with one valve and a gain of about 15dB. Two valves in parallel should give 100W output. 5W to 40W transistors are available, but expensive and delicate.

RELAYS: Use good change over relays with low loss.

When a masthead pre-amp is used, these must also be mounted at the masthead, with protection against the weather. Two coax leads are then run from change-over box on the mast down to the shack.

A "good" relay means one with sufficient isolation between ports (better than 40dB) and low VSWR (less than 1.2 to 1) as well as low loss. Many of the cheaper types have poor isolation and VSWR and the loss, in dissipative terms, is often not the limit. When the relay IS good then the real dissipation causes the power handling limit. The type manufactured by TRANSCO is an example of a good relay.

A high power relay is needed to switch the antenna between the transmitter and receiver. A second low power one is then often used to switch the preamp input between the main relay RX port and a 50ohm load. Then on transmit the pre-amp is terminated in a 50ohm load. The additional isolation from the low power relay also protects the preamp when the main relay isolation drops while it is switching.

The following references in Radcom will provide some more information:

G3JVL Loop Yagi	Jul 76, p525; Sep 78, p783 Jan 79, p41; Jul 79, p636 plus VHF Manual p8.47
Stacking Yagi Antennas	Mar 80, p271
Splitters & Combiners	VHF Manual p8.49
Parabolic Dishes	VHF Manual p8.50 onwards
Alford Slot Antenna	Aug 81, p732
High Efficiency 23cm Dish Feed	Oct 81, p941; Sep 82, p773
Heliax Connectors using standard N-type's	Mar 79, p238
Using RF Pre-amps on 1.3GHz	Jun/Jul 80, p653
High Quality UHF Source	Oct 81, p907
Notes & mods on Valve PA's	Apr 79, p342; Feb 81, p147 Feb 82, p151
Double Slug Tuner for 1.3GHz	Jun 81, p536

1.3GHz BEACONS

GB3AND ZL63B, Ikm E Andover, Hants
50W erp, Stacked Slots, 85m asl, omni, FI

GB3BPO AM77j, I0km E Ipswich, Suffolk
700W erp, Slotted Waveguide, 80m asl, E/W, FI

GB3CLE YM48h, I4km NE Ludlow, Salop
20W erp, 2xI5/I5 yagis, 540m asl, N/SE, FI

GB3DUN ZL08e, 5km SE Dunstable, Beds
2W erp, HB9CV, 263m asl, N, FI

GB3EDN YP05g, 3km S Edinburgh, Lothian
25W erp, 2x corner reflector, II7m asl, NE/NW, FI

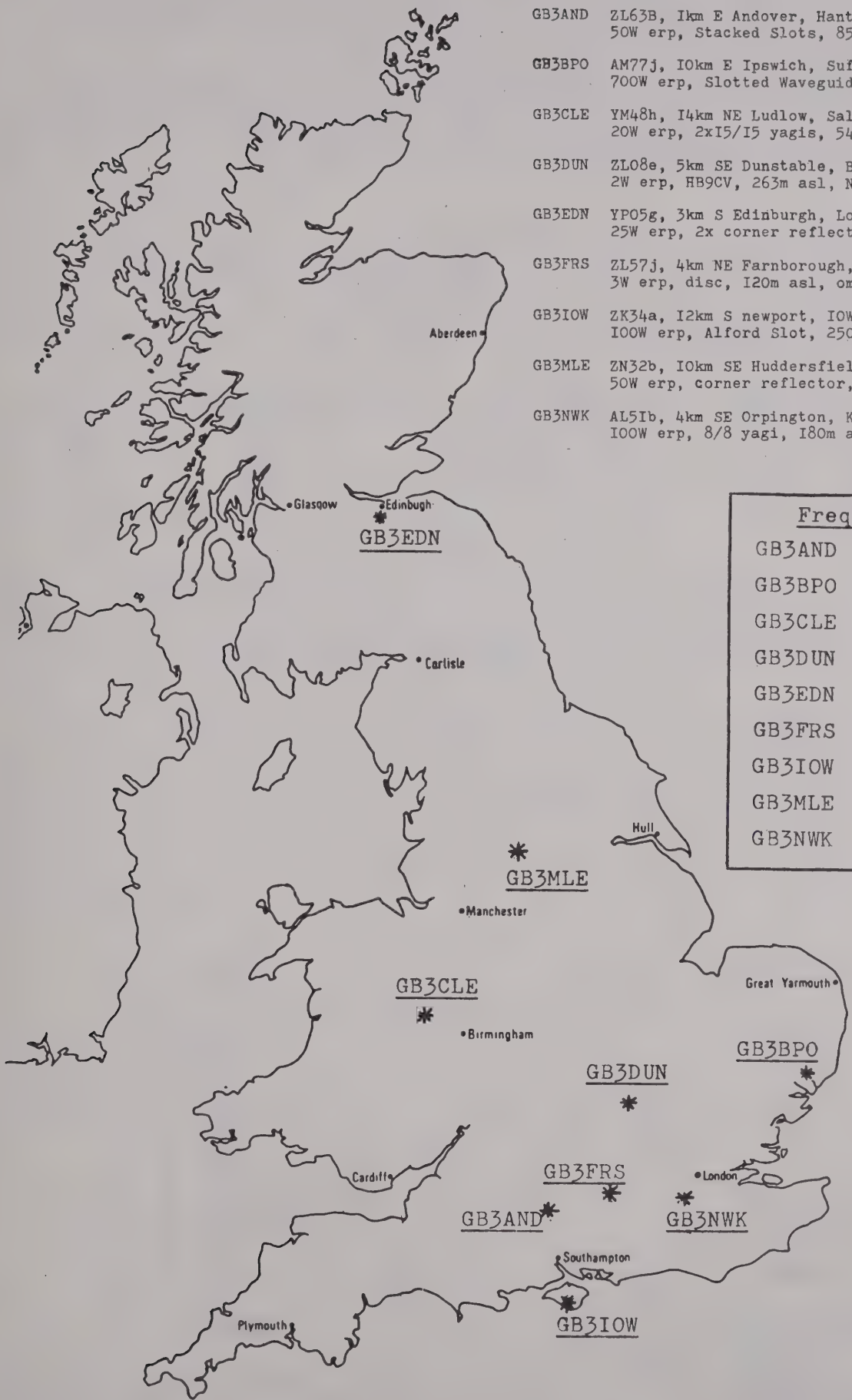
GB3FRS ZL57j, 4km NE Farnborough, Hants
3W erp, disc, I20m asl, omni, F2

GB3IOW ZK34a, I2km S newport, IOW
I00W erp, Alford Slot, 250m asl, omni, FI

GB3MLE ZN32b, I0km SE Huddersfield, W Yorks
50W erp, corner reflector, 600m asl, I60°, FI

GB3NWK AL5Ib, 4km SE Orpington, Kent
I00W erp, 8/8 yagi, I80m asl, WNW, FI

Frequencies	
GB3AND	I296.870
GB3BPO	I296.830
GB3CLE	I296.9I0
GB3DUN	I296.890
GB3EDN	I296.990
GB3FRS	I296.850
GB3IOW	I296.900
GB3MLE	I296.930
GB3NWK	I296.8I0



Alford Slot Antennae for 1.3GHz - G3JVL/G3YGF/G4KNZ

Introduction

The Alford Slot antenna, which has been developed for 1.3GHz by G3JVL, is an easy means of obtaining an omni-directional radiation pattern with horizontal polarisation. The antenna has a gain which depends principally upon its length and is typically 5 to 9 dBi. This is a better performance than other simple omni-directional antennae commonly used such as halos or whips.

It is particularly suitable for a beacon or repeater antenna where an omni pattern is required with as high a gain as possible. In this application it is possible to stack two such antennae end to end (as used at the beacon GB3IOW) and nearly double the gain. With higher path losses on 23cm compared to 2m and 70cm the extra gain makes it particularly useful as a mobile antenna.

Description

The antenna consists of a length of slotted tubing as shown in figures 1 and 2. By suitable design the feedpoint can be either in the centre (fig 1) when the slot is shorted at both ends, or at one end when only the other end is shorted (fig 2). The width and length of the slot, the wall thickness and the diameter of tubing are all related and much experimental work has been done by G3JVL and G3YGF to evolve some working designs, details of which are given below:

Antenna Type	Tube Dimensions	Slot Width	Slot Length
End Fed	31.8mm OD, 20swg wall	4mm	255mm
End Fed	35.0mm OD, 1.1mm wall	8mm	255mm
End Fed	38.1mm OD, 16swg wall	11mm	255mm
Centre Fed	31.8mm OD, 20swg wall	4mm	510mm
Centre Fed	35.0mm OD, 1.1mm wall	8mm	510mm
Centre Fed	38.1mm OD, 16swg wall	11mm	510mm

The dimensions cover three common sizes of tubing available (copper, brass and aluminium materials are all suitable). If they are not followed exactly then some experimentation will be necessary for correct operation. In any case it is advisable to check the field distribution in the slot as explained later.

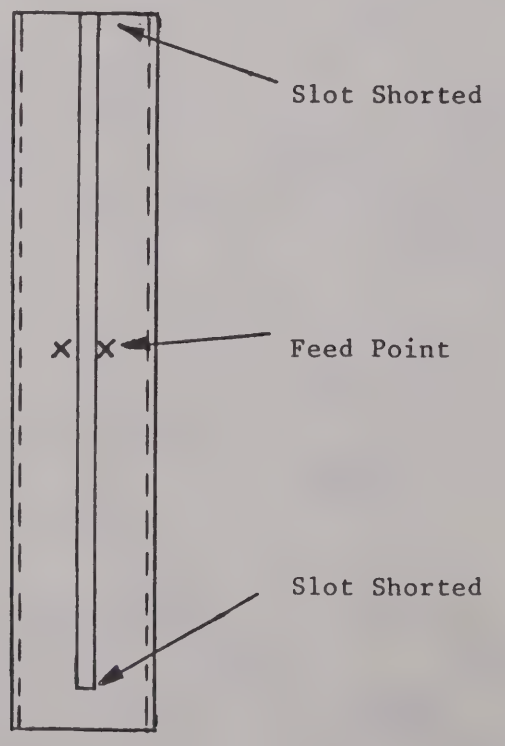


Figure 1

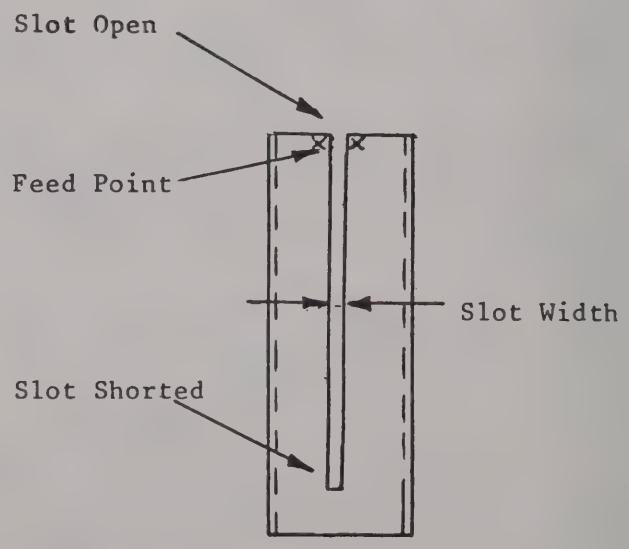


Figure 2

The length of tube beyond the slot is completely uncritical and the same tube could be used both as a mast and as an antenna! This includes the length of tubing above the short in the centre fed version so that either a simple short across the

slot or a disc covering the top can be used, or the tube can be extended upwards in a similar manner to the bottom.

The feed impedance of these antenae is approximately 200 ohm. A convenient method of feeding from 50 ohm coax is to use a 4:1 balun designed by G3JVL which is fabricated from semi-rigid coax, as shown in figure 3. It consists of a piece of 0.141 inch (3.6mm) semi-rigid with two slots cut along opposite sides of the outer. The two leaves formed by the coax outer form a twin wire transmission line which is a quarter wave long, and short circuited at one end. This quarter wave resonator is excited by connecting the coax inner conductor to the end of one of the leaves. The two sides of the semi-rigid a and b are connected to the feed point of the slot (see figs 1 and 2). A convenient method of doing this is to attach small solder tags to the cable so that small screws can then be used to attach the balun assembly to the sides of the slot as shown in figure 4.

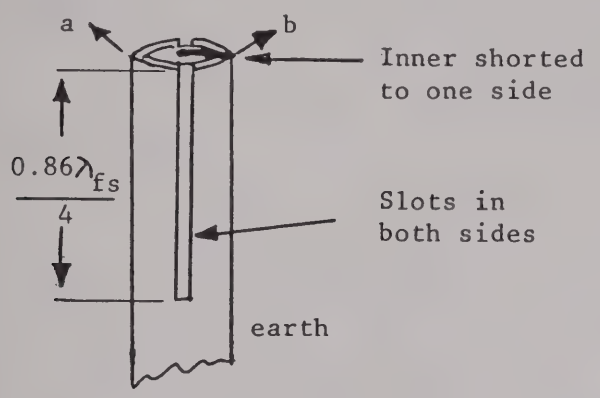


Figure 3

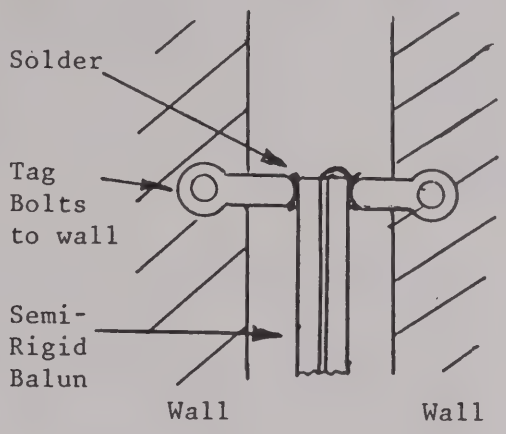


Figure 4

The cable should be bent round after leaving the feed point so that it sits somewhere between the back wall and the centre as it passes down the tube. The exact arrangement is uncritical so long as the cable does not come too close to the slot and upset its operation (apart from the feed point of course). See figure 5.

It is not necessary to connect the cable to the inside of the tube as it passes out of the bottom. However, a convenient method of mounting is to fit a shorting plate of some description across the bottom with an N-type plug or socket in it. The antenna can be mounted entirely by the N-type connector as shown in figure 6. This method is particularly convenient for mobile use when the N-type can be screwed on to a female back to back bulkhead N-type fixed in the roof. This feedthrough in the roof can of course be used for other bands as well. Obviously many other methods of mounting are possible.

Figure 5

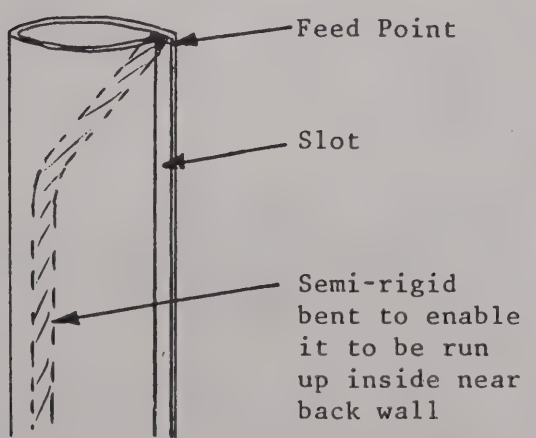
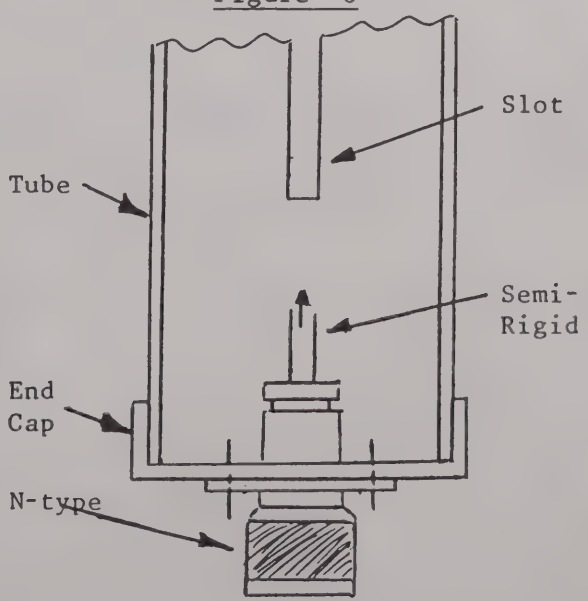


Figure 6



Notes on Construction

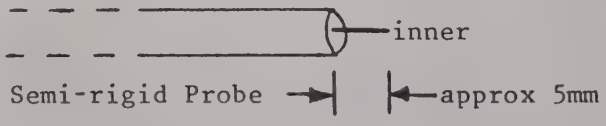
1. The slot in the tubing can be cut with a hacksaw blade and filed to size. For the centre fed version it will be necessary to drill a few holes to start off with.
2. If the tubing used is a plumbing material (eg 35mm copper central heating piping) then other fittings will be available. In particular a pipe blanking cap can be used at the base which will solder or clamp to the tube and in the centre of which an N-type connector can be mounted to bring the coax into the tube from the outside world.
3. The semi-rigid coax for the balun can be held in a vice and bent slightly while the cuts are made. Care should be taken not to cut into the dielectric too much. The leaves should be kept in contact with the PTFE dielectric, and not bent apart at all.
4. At the feed point two holes can be drilled and tapped (eg M2) to fasten the solder tags. Alternatively the tags can be directly soldered to copper or brass tubing and the balun fastened to these later (a blow torch being needed for the first operation, a soldering iron sufficing for the second).
5. The presence of moisture on the inside of the tube will not affect its operation, apart from the balun getting wet, which will introduce a slight loss. However, water will accumulate in the tube and this is not desirable. The slot can be sealed with PTFE adhesive tape and a plastic cap can be used to seal the top in the case of the end fed version. An alternative approach is to enclose the whole assembly in a container such as a sealed length of plastic drainpipe. This method has been used successfully at GB3IOW. In the designs given the velocity factor is approximately 4 and the bandwidth 100MHz at 1.3GHz. The gain achieved for the dimensions given will be about 6dBi for the end fed version and 8dBi for the centre fed version (double slot).

The dimensions are, to a certain extent interdependant. The velocity factor will be increased by decreasing the tube diameter, or by increasing the slot width. The wall thickness also has an effect since it determines the capacitance across the slot so that a thinner walled material will also increase the velocity factor. Thus if a slightly smaller diameter tube was chosen than one of the designs then this could be compensated for by using a slightly narrower slot so that the same velocity factor is achieved. Alternatively the length of the slot could be decreased. The antenna would then operate with a lower velocity factor, but this would give a lower gain. For 1.3GHz antennae the tube diameter should be within the range of those given, any tube much beyond these limits will not operate correctly.

It is important that the operation is checked, particularly if any of the original design parameters are changed. This may be done by feeding the antenna with a signal at various frequencies and looking at the voltage distribution using a power meter, detector or analyser with a small probe to pick up the radiated signal (see figure 7). The probe should be held close to the tube, but not directly in front of the slot (hold it 20 or 30 degrees round from the edge) and moved along its length. The diode current meter described in the microwave newsletter (08/81) would be suitable for this purpose.

The balun works by taking the voltage on the unbalanced 50 ohm line and producing two output voltages relative to earth (the cable outer) which are equal to the input voltage but are 180 degrees out of phase with each other. The balanced load is connected between these two outputs and sees the difference between them, which is twice the 50 ohm voltage. Hence there is a 4:1 step up in impedance. The balun has a comparable bandwidth to the slot, about 10 or 15%. Note that the length of the cuts in the semi-rigid must be an electrical quarter wave long. Since the space between ther inside is PTFE and the space around them outside is air this gives an effective velocity factor of about 0.86. Thus the length is 0.86 times the free space quarter wavelength. If there is a significant gap between the leaves and the PTFE then the velocity factor will be slightly higher.

Figure 7



Summary

The antenna represents a very practical means of realising horizontal polarisation with an omni-directional pattern and high gain on 1.3GHz. The bandwidth is sufficient to cover all of the band so that it would be suitable for any modes including TV. The circularity is very good (ratio of max to min gain) being typically 1 dB. This type of antenna has also been used on other bands successfully - G3JVL has used it on 2m, 70cm and 13cm. For further details contact Mike Walters G3JVL, or Julian Gannaway G3YGF, or the RSGB Microwave Committee.

1.4 1.3GHz Repeater Duplexing Filters

Mike Walters, G3JVL has been performing a lot of calculations recently on waveguide cavity filters similar to those used in the "JVL" 10GHz narrowband tranceiver. One application is for 1.3GHz repeaters to combine the transmitter on 1297MHz and the receiver on 1291MHz into the same antenna.

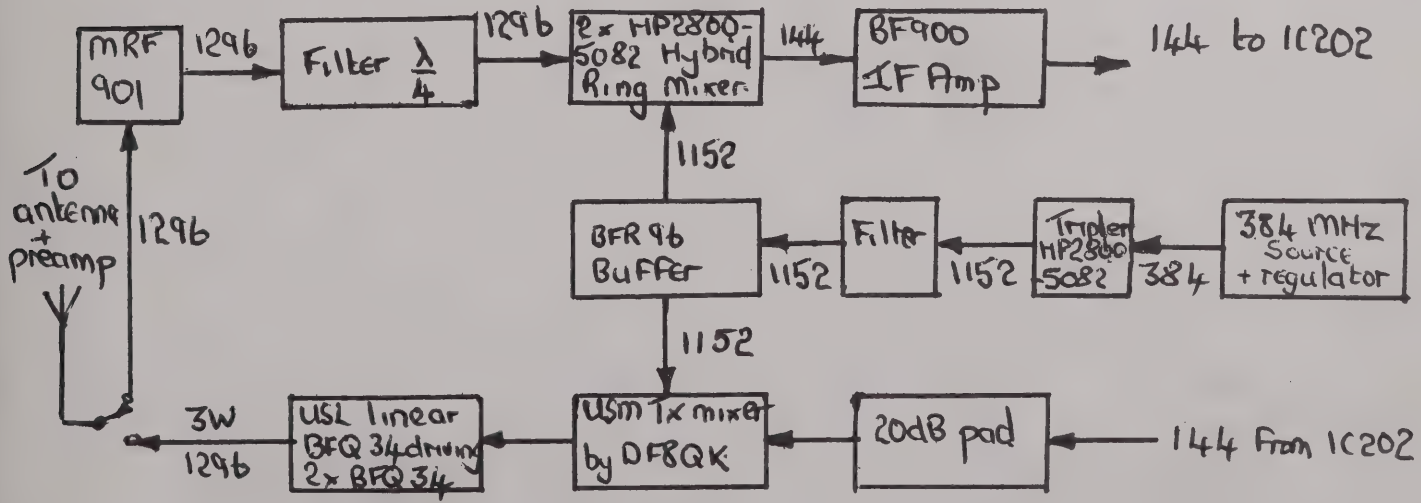
The following design is for 3MHz wide filter on 1297MHz with a rejection of 60dB at 6MHz from the centre frequency using a four pole network. The waveguide used is WG6 of inside dimensions of 6.5 by 3.25 inches (quite big!). Mike has derived the following dimensions:

Cavity Length	=	157.5	161.5	161.5	157.5mm
Waveguide used = WG16		Cavity 1	Cavity 2	Cavity 3	Cavity 4
Iris No	=	1	2	3	4 5
Iris Diameter	=	49.26	23.10	21.48	23.10 49.26mm

The power can be coupled into and out of the filter via transitions to coax, but since two such filters would most likely be used in a repeater, one on the transmit and one on the receive frequency, they could be placed back to back (cf the "JVL" 10GHz transverter) with the antenna connected to a coupling probe or loop in the centre in place of the mixer diode. Anyone contemplating building the filters in such a configuration should contact Mike for further design details. When made of either brass, copper or aluminium, the theoretical loss for such a filter is approx 0.7 dB and the unloaded Q of the cavities is around 20,000. These figures contrast dramatically with the performance obtainable using stripline construction.

1.5 23cm Transverter system

GW4LXO reports now being qrv on 23cm, transverting from 2m. He has 3W to 15/15 ele jaybeam yagi plus a NE645 masthead preamp on receive. For the tx converter he is using the USM and USL bits from SSB Electronics (see diagram) (available from Piper Communications) which he found very easy to build and set up.



GW4LXO 23cm Transverter

Section 2 2.3GHz

2.1 page1

44 Element 2.3GHz LOOP QUAD

G3JVL

For 2320MHz to 2322MHz (New Frequency)

Many people are familiar with the Loop Quad design for 1.3GHz by G3JVL (Feb 1974). A few years later a 2.3GHz design was also produced (Dec 1977) but was not widely published. The dimensions for the 2.3GHz version have been corrected for the new narrowband segment now in use, 2320 to 2322MHz, and are published here.

Boom diameter: 0.5 inches
Boom length: 2 metres approx
Boom material: Aluminium alloy

The element length is measured when unbent, and is the distance between the centre of the two fixing holes. Either 20swg or 22swg aluminium alloy material can be used for the elements; two slightly different lengths are quoted depending upon the size used:

Element	Length for 20swg	Length for 22swg
Reflector plate	3.1" x 2.5"	3.1" x 2.5"
Reflector loop	5.59"	5.58"
Driven element	5.24"	5.23"
Directors 1 to 12	4.69"	4.68"
Directors 13 to 20	4.54"	4.53"
Directors 21 to 36	4.37"	4.36"
Directors 37 to 41	4.27"	4.26"

The reflector plate is fixed 0.25" from the bottom edge in the centre. Element widths are all 0.1875" (3/16") wide.

What is important when marking out the spacings is to keep the total error from one end to the other as small as possible. This error (the RMS error) in the total length is in fact more important than errors in the spacings between individual elements. For example, if each spacing was marked out to be 0.01" too small then the whole antenna would be built 0.44" too small (nearly half an inch). Thus running totals are given along with the spacings between each pair of elements and these should be used to keep down the total error.

Spacings	Running total	Spacings	Running total
RP	0.00"	D20	35.20"
> 1.547"		> 1.990"	
R1	1.55"	D21	37.19"
> 0.717"		> 1.990"	
DE	2.26"	D22	39.18"
> 0.626"		> 1.990"	
D1	2.89"	D23	41.17"
> 0.464"		> 1.990"	
D2	3.36"	D24	43.16"
> 0.995"		> 1.990"	
D3	4.35"	D25	45.15"
> 0.995"		> 1.990"	
D4	5.84"	D26	47.14"
> 0.699"		> 1.990"	
D5	6.04"	D27	49.13"
> 1.291"		> 1.990"	
D6	7.33"	D28	51.12"
> 1.990"		> 1.990"	
D7	9.32"	D29	53.11"
> 1.990"		> 1.990"	
D8	11.31"	D30	55.10"
> 1.990"		> 1.990"	
D9	13.01"	D31	57.09"
> 1.990"		> 1.990"	
D10	15.29"	D32	59.08"
> 1.990"		> 1.990"	
D11	17.29"	D33	61.07"
> 1.990"		> 1.990"	
D12	19.28"	D34	63.06"
> 1.990"		> 1.990"	
D13	21.27"	D35	65.05"
> 1.990"		> 1.990"	

D14		23.26"	D36		67.04"
	> 1.990"			> 1.990"	
D15		25.25"	D37		69.03"
	> 1.990"			> 1.990"	
D16		27.24"	D38		71.02"
	> 1.990"			> 1.990"	
D17		29.23"	D39		73.01"
	> 1.990"			> 1.990"	
D18		31.22"	D40		75.00"
	> 1.990"			> 1.990"	
D19		33.21"	D41		76.99"

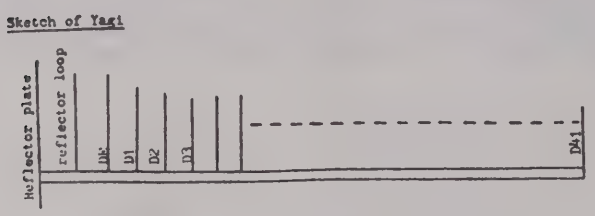
The abbreviations are:
RP - Reflector Plate
R1 - Reflector Loop
DE - Driven Element
D1 - Director 1 etc

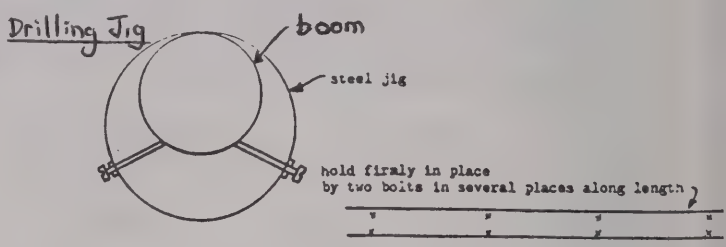
Notes:

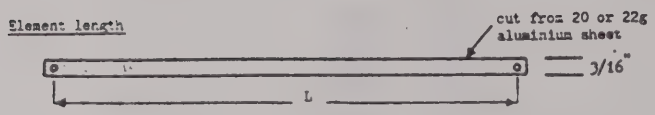
1. If you are making more than one antenna, it is worth making a "jig" out of hollow steel tube. This tube should be bigger than 0.5" inside diameter so that a boom length fits easily inside. Mark out this jig very carefully and drill out holes at each element fixing point. Then insert a boom length, fix in place (see sketch below) and simply continue drilling through each hole. Repeat for the remaining boom lengths. This method means that the accurate marking out has only to be done once.

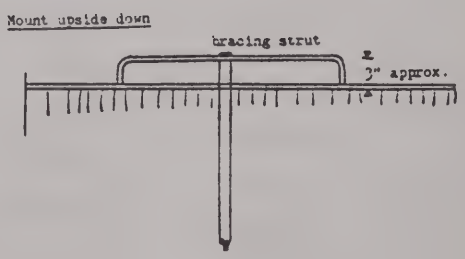
2. After construction either paint the the antenna or varnish it. This will prevent corrosion where the loops are joined onto the beam. Otherwise a gradual loss of gain will be noticed as the aerial corrodes at these points.

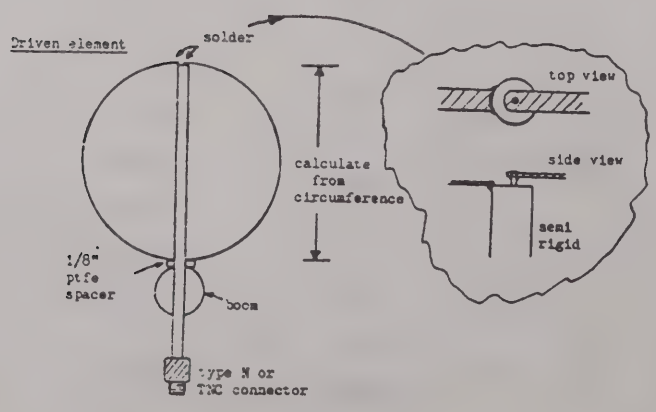
3. It is a good idea to mount the antenna "up-side down" ie with the boom on top. This prevents the elements being squashed when sat upon by large birds.





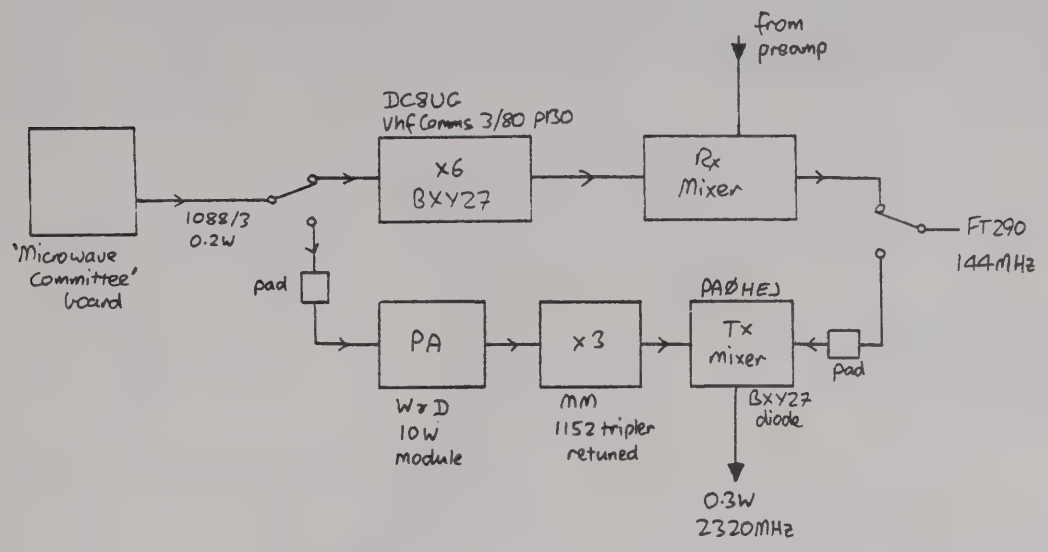






2.2 2.3GHz Transverter System

Dave, G4FRE sends details of his 2.3GHz transverter, which produces 0.3W SSB 'with not a valve in sight'.



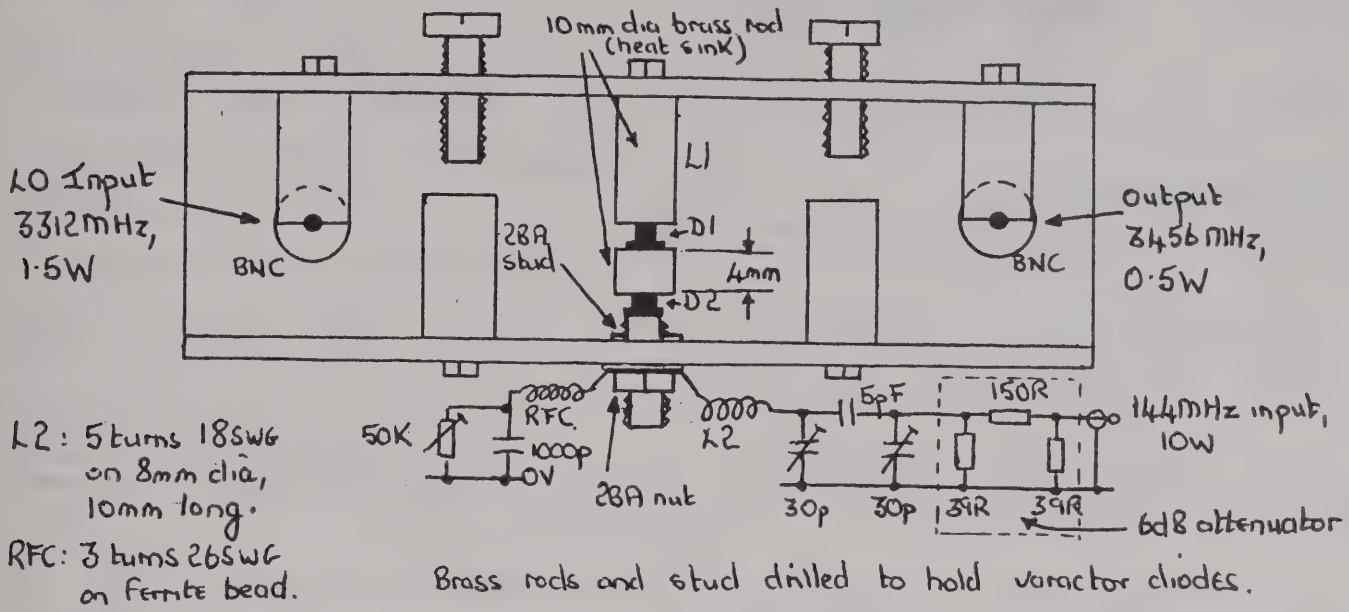
G4FRE 2·3GHz transverter system

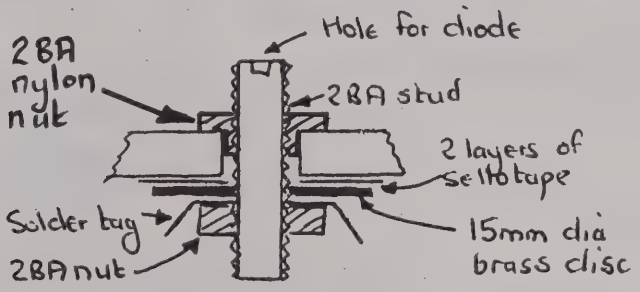
His Rx is the DC0DA interdigital mixer design (VHF Comms 3/78 p154) preceded by a BF981 preamp.

Section 3 3.4GHz

9cm High Level Mixer

The following high level mixer is used by G4MBS on 9cm, producing 0.5W. The tuning elements are taken from the design by DC0DA in VHF Comms P157, 3/1978, except the middle element is changed to solid brass, as is the element between D1 and D2 (for heatsinking). Reference should be made to the above article.





Mounting of brass stud

With 2.5W of 144 MHz and 1.5W of 3312 MHz a single diode (BXY27, 28, 39) produced little more than 100mW but a similar pair in series produced 500mW on 3456 MHz. By using a BXY39 for D1 and BXY28 for D2, Clive managed to obtain as much as 900mW out of the mixer.

The attenuator on the 144 MHz input is essential to isolate the 2m rig from the mixer, and makes tuning up much easier (most rigs will produce 10W so the attenuator would be needed anyway). The trimmers interact and a moderate amount of care is needed when tuning up - Clive suggests tuning up with half power input on 2m, then gradually increasing it to the full level. The 3312 MHz level was measured as 23dB down on the 3456 MHz level.

The local oscillator was generated from an 1104 MHz tripler. Initially an interdigital filter was used between this and the mixer to filter out the 2nd harmonic (2208 MHz), but later dispensed with, with no adverse effects. Care is needed to ensure that noise is not emitted by the diodes and picked up by the receiver if the local oscillator is fed to the tripler continuously - this can be minimised by adjustment of the 50k pot, then readjusting the trimmers for maximum output again.

The 2BA nylon nut is filed down to fit in the hole in the case for about half its length, leaving a lip at one end to prevent it pulling through the hole.

3.2 Waveguide Components

3.2.1 Waveguides & Filters for 3.4GHz

Waveguides for 3.4GHz

There is a wider choice of waveguides for this band, due to the fact that there is a WG11 and WG11A and a WG12 and a WG12A. The WG11 is obsolete, being replaced by WG11A - the latter has a standard aspect ratio of 2.0. WG12 seems to be still used, while WG12A is possibly obsolete; the latter has a rather high aspect ratio of 3.0 and is believed to have been used by the Post Office rather extensively in the past. The choice will no doubt depend upon what can be picked up at rallies etc.

Waveguide No	Dimensions (internal)	Cut Off Freq
WG11	60.25mm x 28.50mm	2.49GHz
WG11A	58.17mm x 29.08mm	2.58GHz
WG12	47.55mm x 22.15mm	3.15GHz
WG12A	50.80mm x 16.92mm	2.95GHz

Waveguide WG10 would also be suitable, though a smaller size is to be preferred usually on weight considerations.

Filter dimensions in WG11A

If WG11A is used instead of WG11 the following dimensions which have been calculated by G8AGN should be used. Lengths are uncorrected for tuning - reduce by 5%, say

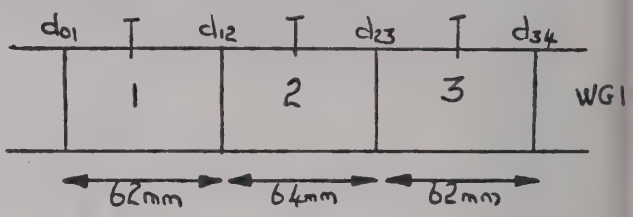
(a) 3312MHz filter, 20MHz Bandwidth:
Iris diameters are 24.2mm, 13.6mm, 13.6mm and 24.2mm.
Cavity Lengths are 68.5mm, 71.2mm, and 68.5mm
(b) 3456MHz filter, Bandwidth 20MHz:
Iris diameters are 22.8mm, 12.5mm, 12.5mm and 22.8mm
Cavity Lengths are 62.2mm, 64.5mm and 62.2mm.

More info will be published when available. Thanks are due to G3JVL and G8AGN, both of whom have written computer programs to do these calculations (contact them direct for info if interested).

Filters for 3.4GHz using WG11

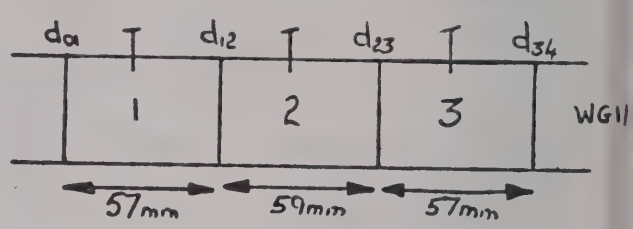
G4KNZ has calculated dimensions for local oscillator and signal filters for 3456MHz (using 144MHz IF). Note that this waveguide is slightly different in dimensions to WG11A (dimensions 58.2 by 29.1mm inside). WG11A is the guide in common usage commercially, which replaced WG11 and if this guide is to be used, a slight correction to the dimensions will be needed.

1. Local Oscillator Filter for 3312MHz, 20MHz wide:



Iris hole sizes $d_{01} = 23mm$ $d_{23} = 12mm$
$d_{12} = 12mm$ $d_{34} = 23mm$

2. Signal Filter for 3456MHz, 20MHz wide:

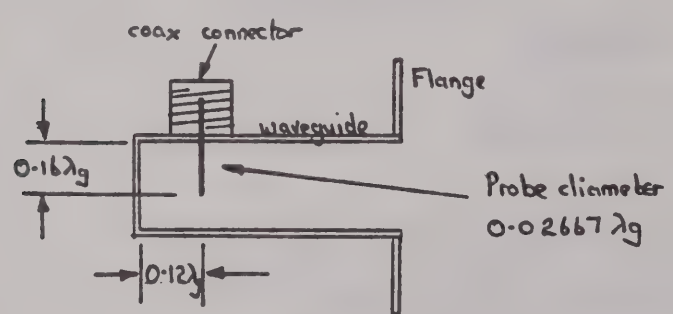


Iris hole sizes $d_{01} = 23mm$ $d_{23} = 11mm$
$d_{12} = 11mm$ $d_{34} = 23mm$

3.2.2. Waveguide Transition for 3.4GHz - G4KNZ

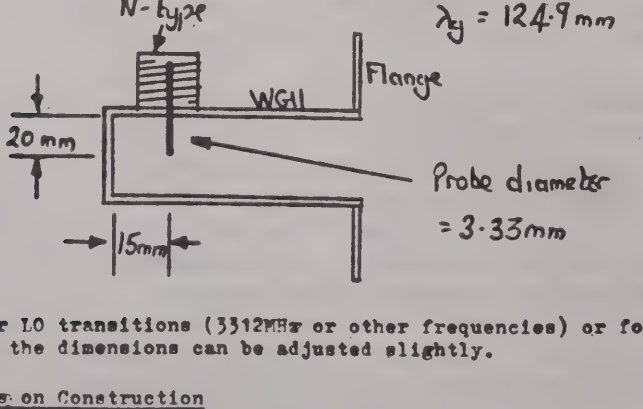
Waveguide Transitions

The following design is for a waveguide to coax transition for use on 3456MHz. WG11 is used but WG11A could be used with little change in dimensions.



The design is based upon some results for 10GHz supplied by G3JVL, when the above dimensions are required (in terms of waveguide wavelength) to produce a transition to a 50 Ohm transmission line.

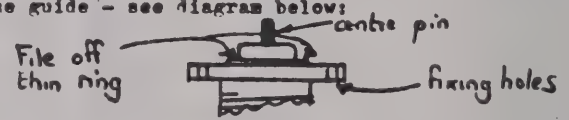
For 3456MHz, using WG11, then the following dimensions are obtained:



For LO transitions (3312MHz or other frequencies) or for WG11A the dimensions can be adjusted slightly.

Notes on Construction

1. A chassis mounting N-type can be used for the coax connector, the one with four holes for bolting to the chassis is preferred. It may be found that part of it has to be filed away for the socket to fit flush on top of the guide - see diagram below:



2. The socket could be soldered to the guide, or held in place by four small bolts, which are cut to the correct len so that the ends end up flush with the inside surface of t waveguide:

3. The outer of semi-rigid (0.141") cable is almost the required size for the probe into the guide. Also, if it is carfully removed from a small piece of cable, it will be found that it is a snug fit over the pin of the N-type cho above. The end of the tubing thus obtained can be filled w a piece of M3 studding which is soldered in place, and the filed flat:

4. The end plate is conveniently made of a piece of brass, several mm thick (2 or 3mm) which is sawn and filed to a tight fit into the size of waveguide in use. It can then b pressed into place and soldered without having to support whilst soldering. A blow torch will be needed for this.

G4KNZ has built a pair of these transitions using WG11 and they seem to work OK, with losses less than 1dB, but n attempt has been made yet to optimise them for minimum los An alternative to making the length between the probe and back plate $0.12\lambda g$ is to make it $0.62\lambda g$. This then means there is sufficient room to include a tuning screw, so tha there is effectively some adjustment of this length possib Making the probe removable - bolting rather than soldering will make it easy to trim the length of the probe if the ultimate in performance is sought.

Section 4 5.7GHz

4.1 5.7GHz Transverter Designs

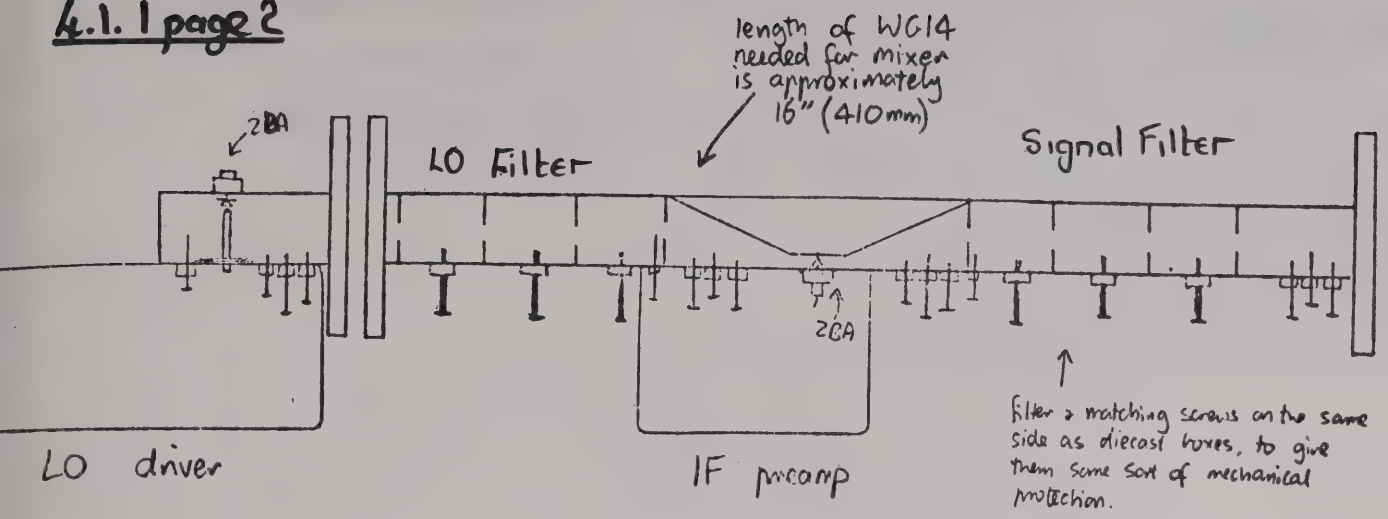
4.1.1 page1

5.7GHz 'JVL Transverter - G4CNV

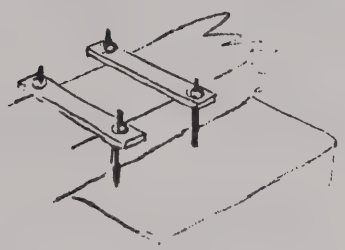
The following design is a scaled-up version of the 10GHz design. The LO and signal filters were designed by G3WDG, the step-recovery multiplier and mixer cavity were scaled by G4CNV from the 10GHz 'JVL. The design incorporates all the features found to be useful in the 10GHz design, so this is closer to a 'second generation' version. The dimensions are shown in the figures. Construction is fairly straightforward, though the following points are worth noting :

1. It is preferable to build the LO source and filter/mixer units as separate assemblies. This makes alignment more straightforward.
2. The diagrams show the LO driver and IF preamp units mounted in diecast boxes strapped to the waveguide. This is a reasonably elegant way of building the transverter, but beware of suspending the transverter directly by its input/output flange, as this puts excessive mechanical strain on the waveguide soldered joints. It is better to use a pair of coax-WG transitions and heliax, or flexiguide (if available) to connect to the antenna.
3. The 347.4MHz source (93.6MHz crystal) delivers 2.5W to the step-recovery multiplier. It consists of the usual "balloon board" (Microwave Committee design) + BGY22 or Wood & Douglas PA. The balloon board should be mounted in its own separate diecast box, which contributes to its stability.
4. The reduced-height waveguide mixer easily justifies the extra effort in building it, and makes the mixer matching screws practically unnecessary In G4CNV's version, within 10 minutes of starting alignment, we had 4mA mixer current. "Remarkably uncritical" was G3YGF's comment on alignment of the 10GHz one.
5. The IF preamp diecast box also contains a low-voltage drop reverse-supply protected regulator (08/80 Newsletter), which also contributes to the excellent frequency stability of the transverter. Both G4KGC and G4CNV have noted that the stability on 5.7GHz seems more than twice as good as on 10GHz. Though this is thought to be due to building the thing properly, having made all the mistakes once on 10GHz!
6. The IF preamp and control circuitry for use with an IC2Ø2 are shown in the diagram. There is little or no difference between this IF preamp and G3WDG's circuit (Radcom, April 1980) - it was simply a matter of availability of transistors.
7. Assembly follows usual waveguide construction practice, jigging tuning and matching screw nuts in place with chrome-plated screws, and soldering everything in one go. The one noticeable difference is that much more heat is needed than for WG16, so use a hotplate on the kitchen stove (made from scrap aluminium sheets), heat everything to just below solder melting-point, and solder using a Ronson blowtorch or soldering iron. Use a good flux (Bakers fluid or similar), and, of course, make sure everything is properly clean to start with.
8. Unfortunately, the procedure recommended for alignment of the 10GHz version is unlikely to be useful due to lack of 5.7GHz wideband components (Gunn osc. etc). However, anyone who has gone through the exercise of aligning a 10GHz JVL will have some idea of what to look for. We cannot yet suggest a foolproof method, but if anyone has any ideas, we'd be most interested to hear them. In the meantime, it will probably be necessary to seek the help of someone with access to professional test equipment, or use the facilities available at Microwave Round Tables.

For any further information or advice, don't hesitate to contact G4CNV, G3WDG, G3JVL or any of the Microwave Committee.


length of WG14 needed for mixer is approximately 16" (410mm)
2BA
LO Filter
Signal Filter
2BA
LO driver
IF preamp
filter & matching screws on the same side as diecast boxes, to give them some sort of mechanical protection.

diecast boxes bolted to W/G by 30mm long M3 bolts and 2 off 2" x 5/16" x 1/8" brass straps.

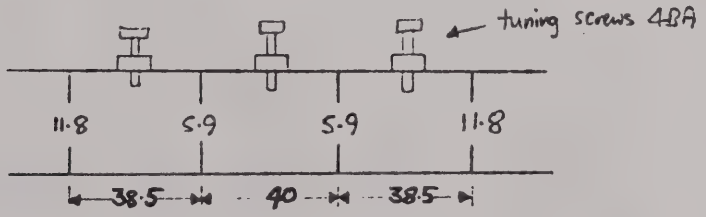


filter tuning screws are 4BA / M3
matching screws are 4BA / M3
intra-filter mixer matching screws are 8BA/M2
mixer matching screws are probably unnecessary

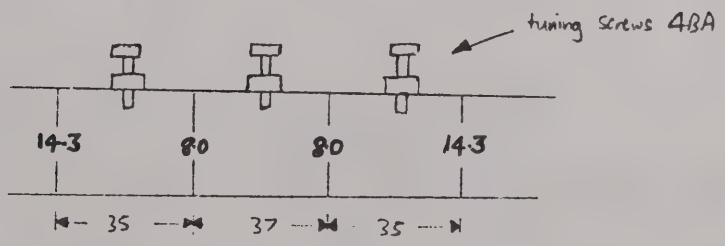
G4CNV

WG14 Filter dimensions (G3WDG)

LO filter 5616MHz, 20MHz b/w, $\epsilon = 0.3333$



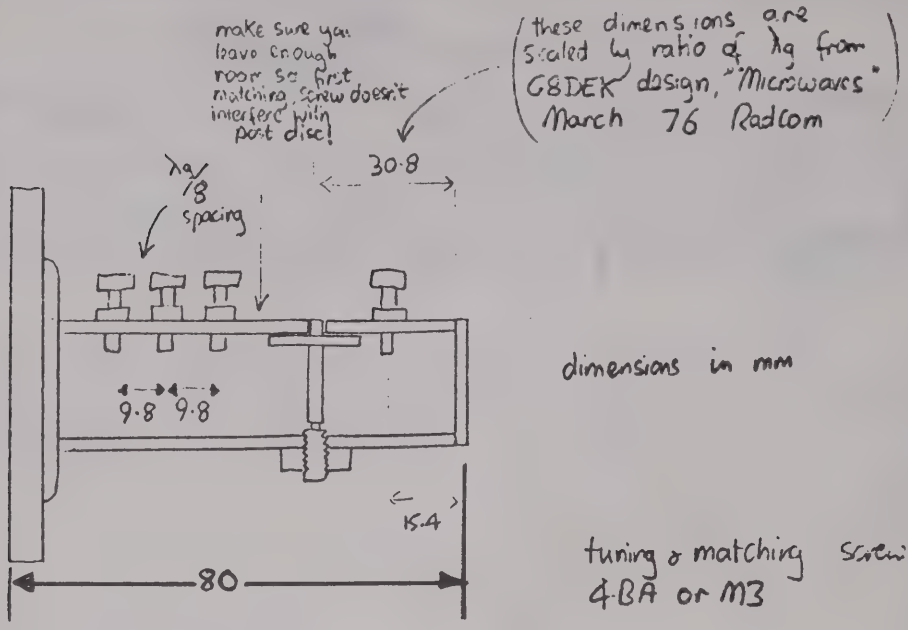
SIG filter 5760MHz, 60MHz b/w, $\epsilon = 0.1$



iris thickness 0.6mm

all dimensions in mm
use Copper WG14, or filter insertion losses will be excessive.

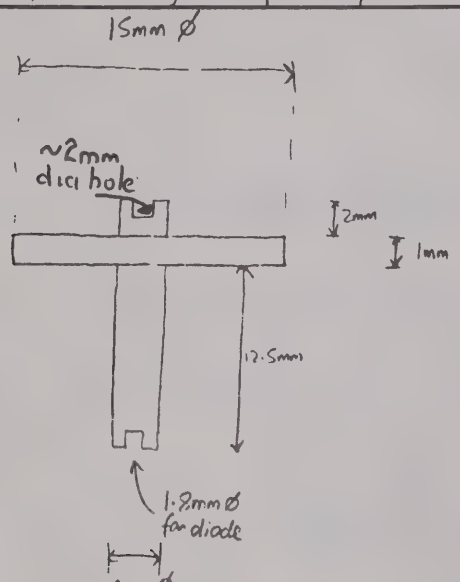
5.7 GHz step-recovery diode multiplier dimensions


make sure you leave enough room so first matching screw doesn't interfere with post disc!
(these dimensions are scaled by ratio of λg from G8DEK design, "Microwaves" March 76 Radcom
λg/8 spacing
30·8
9·8
9·8
15·4
80

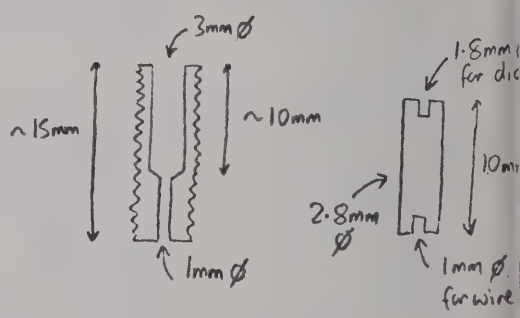
dimensions in mm

tuning & matching screws 4-BA or M3

5.7 GHz step-recovery multiplier post dimensions



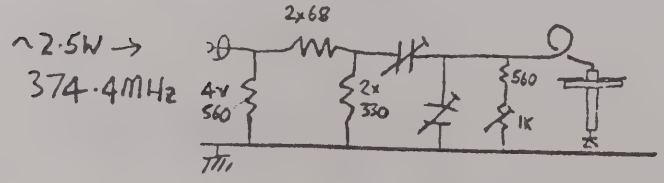
Mixer post dimensions



2BA studding

wrap with sellotape, & use ptfe pressure pad as with 10 GHz design

374.4 MHz i/p matching exactly as for 10 GHz design ("Microwaves" March 76 Radcom)

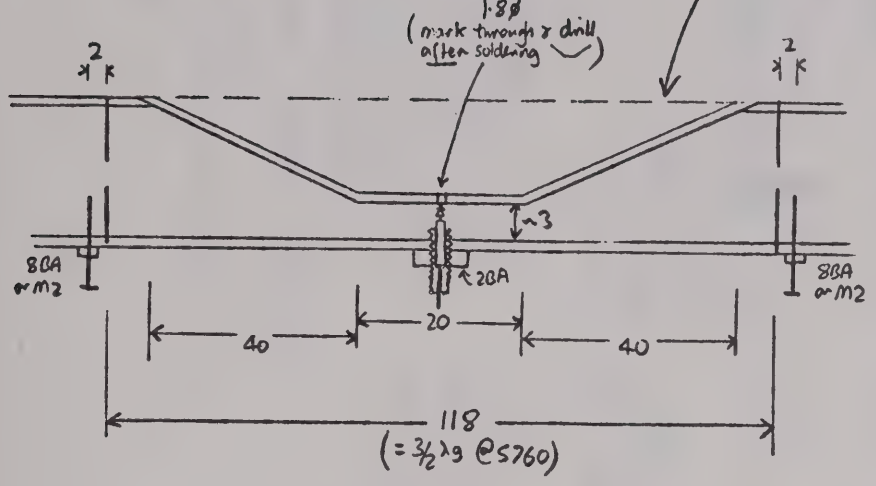


Cs are Mullard 1·8–10pF

4.1.1 page 4

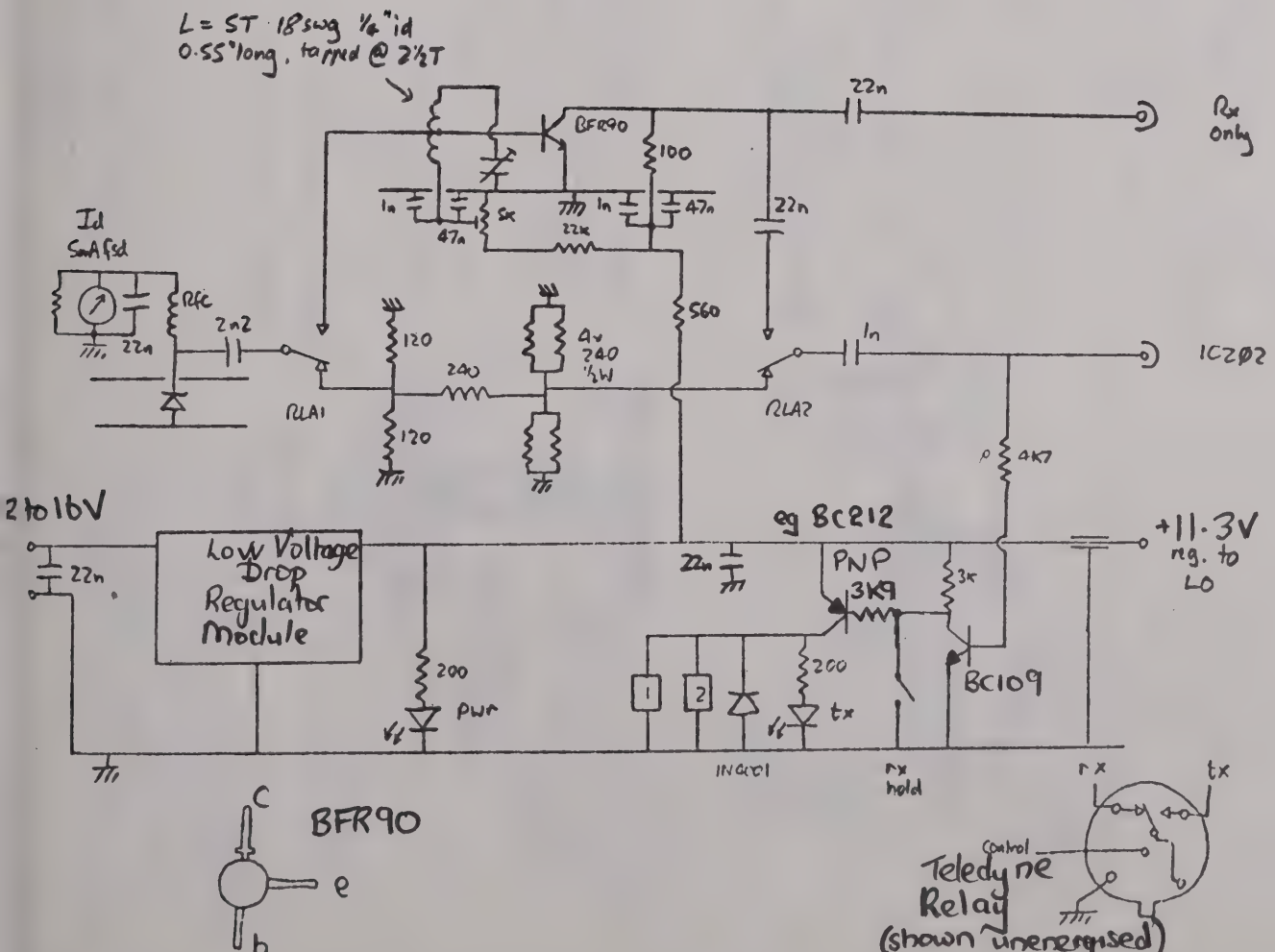
5,760 MHz reduced-height guide mixer
(scaled from G3JVL design, 01/81 Newsletter)

Broad face of W/G removed, but leave walls intact right to the top for rigidity. Bend ⌣ from 18 or 16g brass or copper, lay in place, & solder at the same time as everything else. Nuts jigged with chrome-plated screws in tapped holes in W/G. Soldering best done on kitchen stove with big hotplate, heated to just below solder MP, and use Ronson torch for each joint in turn. Make sure everything's clean to start with, & use good flux (Bakers' fluid)



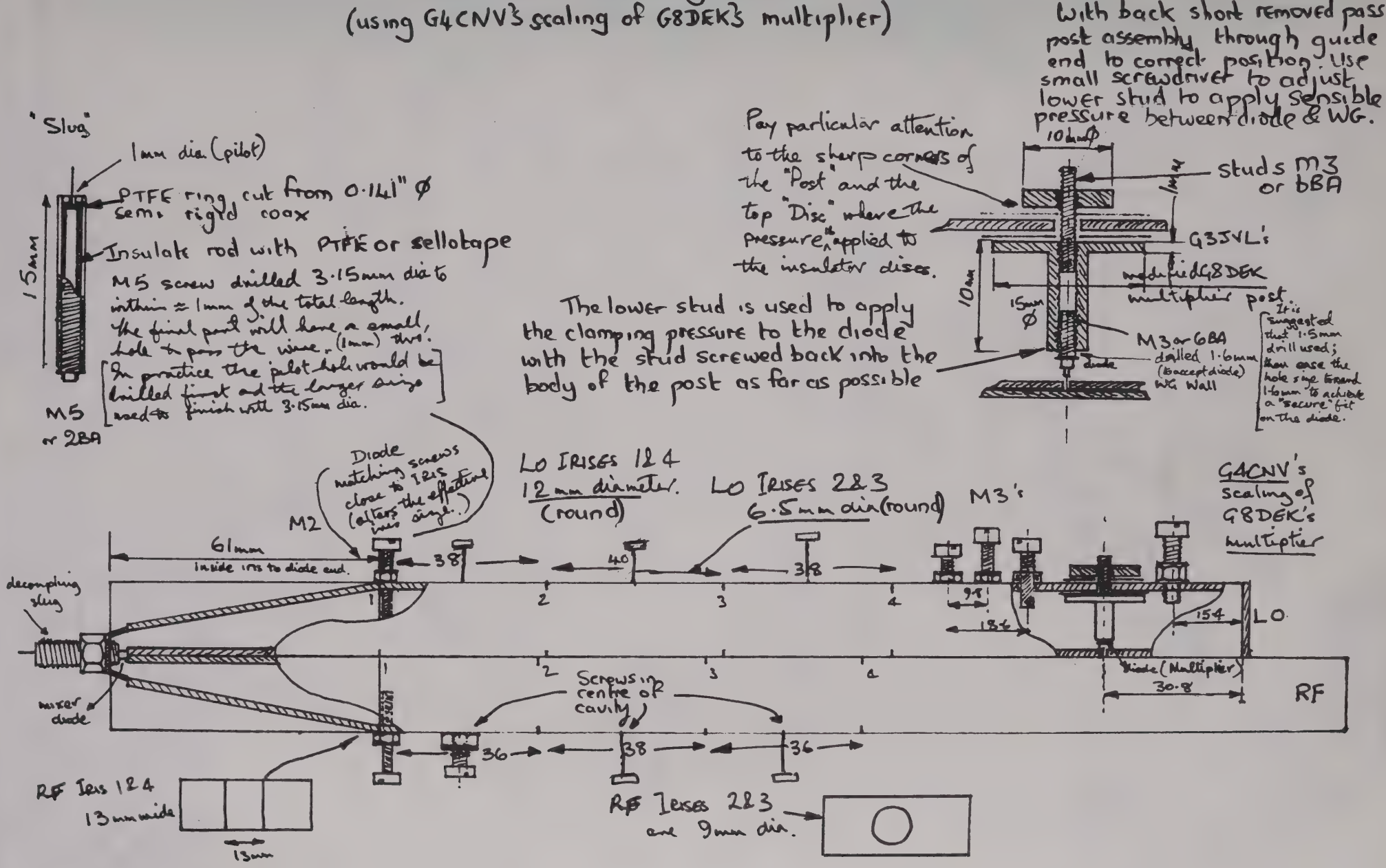
all dimensions in mm

G4CNV



Mixer for 5·7GHz by G3JVL

(using G4CNV's scaling of G8DEK's multiplier)


"Slug"
1mm dia (pilot)
PTFE ring cut from 0·141" Ø semi rigid coax
Insulate rod with PTFE or sellotape
15mm
M5 screw drilled 3·15mm dia to within ≈ 1mm of the total length. The final part will have a small, (1mm) dia. hole to pass the wire.
In practice the pilot hole would be drilled first and the larger drill used to finish with 3·15mm dia.
M5 or 2BA
With back short removed pass post assembly through guide end to correct position. Use small screwdriver to adjust lower stud to apply sensible pressure between diode & WG.
Pay particular attention to the sharp corners of the "Post" and the top "Disc" where the pressure is applied to the insulator discs.
10mm
Studs M3 or 6BA
G3JVL's modified G8DEK multiplier post.
10mm
15mm Ø
M3 or 6BA drilled 1·6mm (to accept diode) thro' Wall
Diode
It is suggested that 1·5mm drill used; then ease the hole size to and 1·6mm to achieve a "secure" fit on the diode.
The lower stud is used to apply the clamping pressure to the diode with the stud screwed back into the body of the post as far as possible
Diode matching screws close to IRIS (alters the effective iris gap.)
M2
LO IRISES 1 & 4 12mm diameter. (round)
LO IRISES 2 & 3 6·5mm dia (round)
M3's
G4CNV's scaling of G8DEK's multiplier
61mm inside iris to diode end.
38
40
38
9·5
18·6
15·4
LO
decoupling slug
mixer diode
1
2
3
4
Diode (Multiplier)
30·8
RF
Screws in centre of cavity
36
38
36
RF Irises 1 & 4 13mm wide
13mm
RF Irises 2 & 3 are 9mm dia.

4.1.2 "Folded G3JVL Mixer for 5·7GHz

Section 5 10GHz General

5.1 Introduction

5.1.1 Making a Start on 10GHz - G4KGC

The following is a list of references from Radcom of information for getting started on 10GHz. Some of the articles can also be found in the microwave chapter of the VHF/UHF Manual.

Construction:	Throughline 10GHz tranceiver	Jun 78, p492
	G8APP Gunn Oscillator	Feb 76, p123
	Directional coupler tranceiver	Oct 73, p692
	IF Preamplifiers	VHF Manual (3rd ed) p8.37
	G3JVL narrowband tranceiver	Jan 79, p41; Apr 79, p342 Apr 80, p372; Feb 81, p146
	Self calibrating wavemeter	Feb 73, p106
	High Q wavemeter	Apr 73, p261
	Simple crytal controlled signal source	May 76, p352
	Horn design	Feb 72, p81; Apr 72, p229
	Dustbin-lid dish	Mar 76, p194
	G4ALN dish feed	Oct 76, p757
	C4COM receiver alignment aid	Jan 76, p36; May 79, p424 Sep 79, p832; Oct 79, p934 Feb 80, p155; Mar 80, p270
	A simple method of microwave frequency measurement	Jan 82, p34
Operating:	Microwave path checking	Feb 78, p122
	Path plotting	Mar 81, p241
	Performance of equipment over line of sight paths	May 78, p413
	Calculating bearings and distances from NGR's	Sep 80, p909
	"Take to the Hills"	Jul 72, p436
	Aligning antennas accurately	Nov 80, p1164
	Sighting Compass	Feb 81, p147
	Polarization	Aug 78, p698
	Superrefraction	Jun 77, p453; Oct 78, p867 Nov 78, p951; Oct 82, p873
	Enhancements	Nov 79, p1044
	Tropo Openings	Mar 80, p270; Apr 80, p382
	Fixed operation using passive reflectors	Jan 81, p47

Most components for 10GHz construction can be obtained at various trade stands at rallies and amateur trade exhibitions. Gunn diodes are available from J. Birkett. Waveguide 16 flanges and waveguide plus step recovery and mixer diodes needed for a G3JVL narrowband transverter can be obtained from J. N. Gannaway, G3YGF at Dean Hill Barn, Dean Hill, Salisbury, Wilts. SP51HJ. Silverstone Electronics stock flanges for WG16, Gunn oscillators and mixers. The address is 78 High Street, Whittlebury, Towcester, Northants. NN12 8XJ. Eighteen inch dishes are available from Practical Wireless, although the f/d of these is a little small at 0.28. Small metal sections (brass, copper etc) can often be obtained from model shops. A good mail order firm for metal sheet, metal rod and tube, nuts and bolts etc is K. R. Whiston Ltd., New Mills, Stockport, SK12 4PT.

Waveguide and flanges for construction of tranceivers etc can often be obtained cheaply by buying surplus lengths of guide with flanges on. If it is necessary to remove the flanges, the method of joining them to the guide should be examined - brazed flanges will prove impossible to remove, soft soldered ones easy (solder easily scratchable) whilst silver soldered joints can be separated with a little care and a good blow torch.

Other sources of information, besides Radcom are:

1. Microwave Newsletter - Ten issues a year are produced by the RSGB microwave committee and distributed by RSGB HQ. It contains a mixture of operating news and technical items. Six of the issues are timed to arrive just before each of the 10GHz cumulatives during the summer. Contact one of the editors - G3YGF, G4CNV or G4KNZ for more details. Subscriptions should be sent to RSGB HQ clearly marked "Microwave Newsletter".
2. VHF Communications - This is a German magazine, published quarterly, but an English version is also produced. Contact SOTA Communications Systems Ltd., 26 Childwall Lane, Bowring Park, Liverpool L14 6TX. It is entirely technical, about half the information on microwaves. Back issues are available, best bought at a rally.

5.1.2 page1

An Introduction to 10GHz Operating in the Cumulatives

G4KNZ

Introduction

A good time to start working on 10GHz is during one of the cumulatives contests when there are plenty of stations try with. They run from 10am until 8pm but most people are around in the afternoon. Compared to a contest on the VHF/UHF bands they are much more relaxed - it would be true to say that most of people treat the day as an activity period rather than a contest. For the rules see April Radcom.

When you first go out on the band you will probably be using simple gear with Gunn oscillators and wideband FM modulation. It may be entirely home built or use useful components such as retuned intruder alarms and doppler modules. This will be good experience before trying some of the more exotic techniques such as ssb.

Choosing a Site

QSO's on 10GHz are very different to those on 2m or 70cm. The difference basically is that propagation works over line of sight paths and you use very narrow beamwidth aerials.

With simple equipment from good sites, line of sight paths up to several hundred kilometres can be worked. Typical paths are 10km to 100km though, because of the geography of the country.

It is because the capability of simple equipment is limited virtually to line of sight paths that most operation on this band is between portable stations. Non line of sight paths are almost impossible with simple equipment. There are exceptions such as the signal enhancements due to ducting which occur over the sea with the right weather conditions. Paths across the English Channel can be worked with low power wideband equipment despite the path being obstructed by many 100's of feet of sea.

You should look for a site which is line of site to several other stations you know will be going out. To find out where other people will be, look at the map in April Radcom or the activity list in this newsletter.

Because it is important to have a line of sight path, trees and bushes stop the path from working. You should make sure your view towards the other station is clear - this point cannot be overstressed. The best hilltops are clear in most directions. Otherwise you may have to move about the site between contacts to work in various directions. Remember when looking at a map that trees may be tens of feet higher than the ground. It is worth visiting the site before the contest.

You will find that several separate stations often operate from the same site. The major difficulty is interference between the talkback stations, but G3YGF and G4KNZ have found (as have others) that at least two stations can operate succesfully from the same site. It does begin to get difficult, however, with 4 or 5 stations co-sited.

Arranging Contacts

If you turn out on the top of a hill and call CQ on 10GHz you will almost certainly not have a single contact. Because beamwidths are so narrow on 10GHz (sometimes as low as 2 degrees) the chances of two stations pointing at each other and being on the same frequency at the same time are remote.

Virtually all tests are pre-arranged using a lower frequency for "talkback". The talkback system must first be able to provide a strong enough signal to attract the attention of anyone else even if they are not beaming at you. If it can do this then it will also provide a reasonable link when both beams are aligned.

Usually 2m is used for talkback. Wider beam widths are used on this band and frequencies are known accurately. Thus the chances of you hearing stations you are looking for is much higher.

Simple hand held rigs and low gain antennas (eg whip or dipole) are adequate for short paths of a few tens of kilometres. For longer paths a more substantial talkback station is required eg a small beam (3 to 8 elements) at 10 or 20ft and around 10W of RF power. Almost everyone uses horizontal polarisation and people with vertical whips are at a disadvantage.

For the last few years 144.330MHz has become a standard microwave calling frequency during a cumulative contest. Unfortunately 144.330MHz is a bit close to the normal calling frequency of 144.300MHz, and this results in QRM. For this reason 144.175MHz has been used last year as well. Thus you

should listen or call around these frequencies during a microwave cumulative contest.

So that weak stations can be heard on these calling channels, people generally QSY after establishing a contact. The regions 144.150 to 144.200MHz or 144.350 to 144.400MHz are most commonly used. If you don't hear anyone on the calling frequencies, try tuning across these parts of the band as people often spend a long time there.

Pointing the Aerial

When you have contacted another station to work using the talkback, you have to point the 10GHz aerial in the right direction.

On VHF/UHF this is usually done by guess work and on HF most antennas are not directional so the problem does not arise. Unless you can see the other station you need to calculate the bearing or draw a line on a map. If the National Grid References of both sites are known calculation is easy. See issue 05/82 of the newsletter for more information. Usually the other station will be happy to calculate the bearing if you have any problems.

Starting out you will probably use a low gain antenna, maybe a horn or small dish, with a beamwidth of 10 to 20 degrees. This will not be too difficult to set up.

Larger dishes may have a beamwidth of only a few degrees and unless pointed in the right direction will have less gain than open waveguide. For a beginner they are something of a mixed blessing and not worth the bother. For example, a 4ft dish, which has a gain of 39dBi, has a 3dB beamwidth of only 2 degrees. The antenna must be aligned to an accuracy of its beamwidth, which in this case means to plus or minus 1 degree.

You can set the beam direction quite accurately with an ordinary compass. For a small horn or dish, judge by eye that it is pointing in the same direction as the compass. A more accurate method involves looking across the face of the dish using a sighting compass. This type of compass allows you to see both the compass needle and the object at the same time.

It goes without saying that the aerial must be mounted firmly so it stays in the direction you point it. The equipment is often mounted on a tripod or it can be put on the roof of a car. Once you can actually hear something on 10GHz from the other station adjust the antenna for maximum signal.

General

In addition to the normal logbook, it is worth keeping a note of the NGR's and bearings from your site to other stations. A record of this information in a separate book will save you many hours - you shouldn't have to repeat the bearing calculations each time.

The Ordnance Survey 1:50,000 maps are very useful for finding good sites - these are the ones with a pink cover, scale about 1 inch to the mile. The O.S. quarter inch maps are also useful for getting an idea of activity and drawing lines. We recommend you use these rather than some of the cheaper "touring" maps.

Have the 10GHz and the talkback near to each other. This is so you can keep the other station informed of what is happening and so he can hear his own microwave signals for peaking up his aerial.

Bring a large polythene bag with you so that when it rains the gear stays dry and working - in showery weather we always hear of a few stations getting caught out and having to pack up.

Sometimes a contact is completed in five or ten minutes, but it is not unknown for a QSO to take up to several hours to complete! A lot of time is often wasted by trivial but unavoidable problems while you get familiar with techniques at these frequencies. For example, with the current level of activity it is frequently possible to hear more than one signal on the band and sometimes it is difficult to distinguish between them.

However, you will usually find other operators patient and willing to persevere with a difficult contact. The whole point about these events is that they are to encourage activity and are a complete contrast to a VHF contest

Relative Performance of AM/FM/SSB - G3YGF

While signals can be quite strong over line of sight paths, it is often the case that the signals sound very strong, but in fact there may only be 10 dB in hand. This is a consequence of the wideband FM modulation system, where the audio S/N can change very rapidly for small changes in the input S/N.

Strong signals, eg an S9 report would mean that the signal was 40 to 50 dB over the noise level (at 6 dB per "S" point). If this were the case, then

1. They would be heard equally well over a range of beam headings, as the average sidelobe level on the dish may only be 15 to 30 dB down on the main beam.
2. They would also be heard well on open waveguide - typical gain for a 2' dish might be 33 dBi and for open guide 6 dBi.

In both of these cases there would still be about 20 dB in hand. Signals this strong are not very common, and although the audio S/N may be 40 to 50 dB, the signal will usually not be audible unless the dish is pointed in the right direction, implying that there is only 10 to 20 dB in hand.

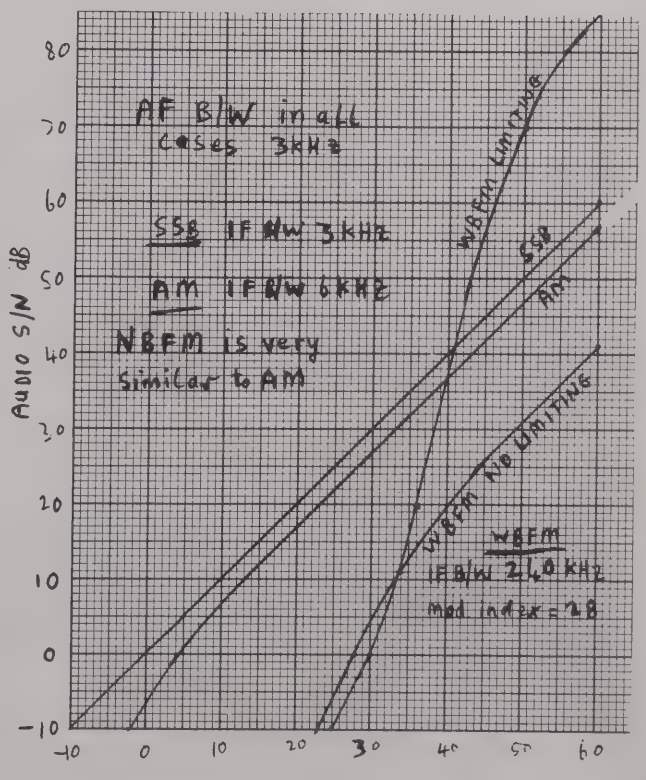
We are not usually concerned with having a very high audio S/N, 30 dB is probably quite adequate. What we want to know is how much more path loss we could tolerate before losing the signals, either by lengthening the path, or allowing more of an obstruction in it. These will all affect the IF S/N, and a very high audio S/N does not mean that we can accept that number of dB extra path loss before signals vanish. The reason for this can be seen in the figure. At low S/N's, (above 0 dB) the audio S/N increases at the rate of 4 dB/dB of input signal. So an increase in signal of 10 dB will take a signal from being barely detectable (0 dB) to apparently S9 (40 dB audio S/N). In fact, for very strong signals - a true S9 wideband FM will give a higher audio S/N than SSB in 3 kHz bandwidth, but for weaker signals SSB or other narrowband modes are definitely superior.

The WB FM system sacrifices weak signal performance for better performance with strong signals. This is mainly due to the limiting action in the detector, and if no limiter is used, the weak signal performance improves by several dB. It should be possible to operate below limiting by reducing the IF preamp gain, or placing an attenuator between the preamp and main IF strip.

In order to obtain maximum range, or contacts over obstructed paths, we will probably be working with weak signals, and it is in these cases that the superiority of SSB, or narrowband modes will become obvious.

Received signal levels can therefore only be properly measured by either a calibrated S meter or by inserting a calibrated attenuator before the limiting IF strip - either in waveguide or between the IF preamp and limiting IF strip. The attenuation can then be increased until the signal is only just detectable, and the level expressed as a number of dB over noise - dBN. This is not an absolute unit, as it depends on the receiver sensitivity, but it is very meaningful, and constant for one receiver, and the noise figure can be measured at some later time.

Graph showing the performance of various modulation methods against signal level:



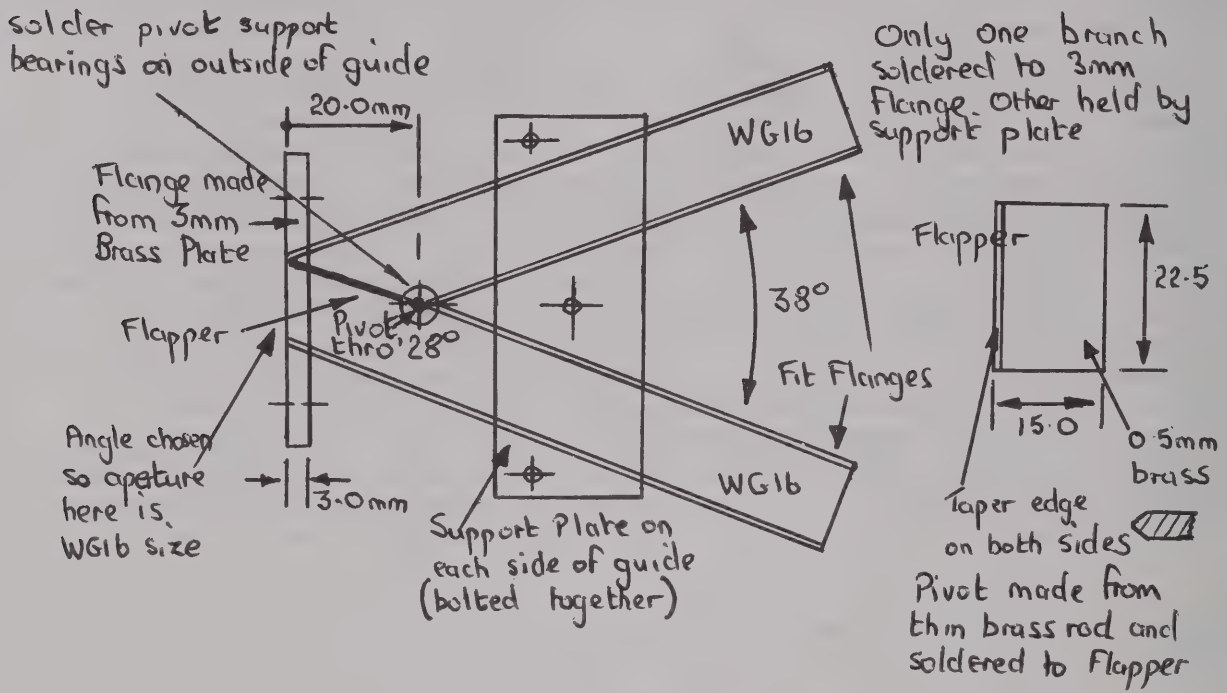
I.F. Carrier to noise ratio (dB) in a constant bandwidth. i.e relative carrier power.

5.3 Components

5.3.1 Waveguide Switch

In VHF Comms, 1/81, there is a design for a waveguide switch as part of a 10GHz ssb system described there. Cliff, G8SHF has built a switch based on this design and found it very satisfactory. For anyone who has not seen this article, the essential details of Cliff's switch are shown here.

Two pieces of waveguide are butted together to form a V-shape which is then joined to a flange. A flapper is then placed at the intersection as shown in the diagram. This is allowed to pivot so enabling either of the waveguide branches to be selected from a common input, and this can be done manually or with an electric motor. In particular Steve, G4AUC suggest using a motor used to operate model railway points, such as the SM3. Cliff has used a Ledex switch coil, which gives about the right amount of travel, 28 degrees, but is rather heavy on current.



5.3.2 Flanges

Joining Round Flanges Together

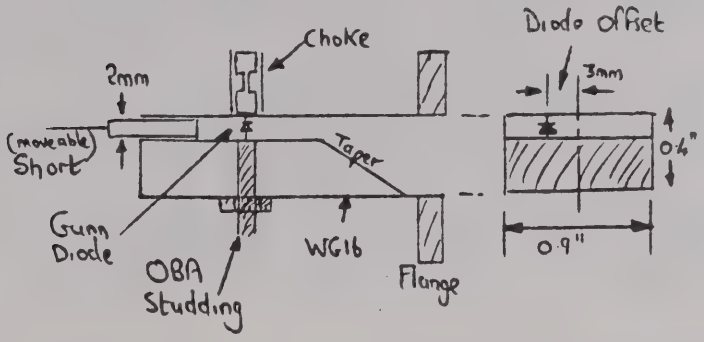
G4MBS says he could do with an extra pair of hands when doing up a pair of WG16 round flanges (particularly when balanced on the top of a ladder)! His answer is to smear a little superglue on the locating ring and then fit it all the way onto one flange. Then it will remain in position for the other flange to be pushed in against the first. This makes doing up the securing ring much easier.

Waveguide and flanges for construction of tranceivers etc can often be obtained cheaply by buying surplus lengths of guide with flanges on. If it is necessary to remove the flanges, the method of joining them to the guide should be examined - brazed flanges will prove impossible to remove, soft soldered ones easy (solder easily scratchable) whilst silver soldered joints can be separated with a little care and a good blow torch.

5.4

10GHz Gunn Reflection Amplifier

Barry, G8AGN has been experimenting with Gunn diodes as amplifiers and suggests the following configuration might work:



He suggests biasing the diode to about 3x the threshold voltage (watching diode dissipation). The max output would depend on the diode's performance as an oscillator and gain would be of order 10dB, with a bandwidth of perhaps 10%. The diode is offset from the centre of the waveguide as shown. Anyone wishing to experiment could obtain more info from Barry, and if it works we will print more details.

5.5 Aerials

5.5.1

Horn Antennas from Oilcans - G3YGF

Small horns are very useful general purpose aerials at 10GHz for short range tests, signal sources etc. They are extremely easy to construct as their dimensions are not critical and they will provide a good match over the whole band. One can easily be built in an evening.

Empty 1 gallon oil cans are an ideal source of tinplate for this purpose. Select cans with no rust visible either inside or outside. Cut off the top and bottom of the can with tin snips or a hacksaw, then cut along the length of each of the four corners and flatten out the four pieces of tinplate thus obtained.

Although a horn of any gain could be built, three values suggest themselves from the sizes of tinplate available: a 16 x 16cm horn of 23dBi gain requiring two cans (it uses four of the larger sheets); an 8 x 16cm horn of 20dBi gain using two of the larger and two of the smaller sheets to give a horn with a similar aspect ratio to that of the waveguide; or an 8 x 8cm horn of 17dBi gain made by cutting dowm two of the larger sheets (or using four smaller sheets) which need only to be about half the length.

Remove any sharp edges or slivers of metal, then wash and degrease the sheets. It is not necessary to remove most of the paint, only where you want to solder. This can be done with wire wool or paint stripper.

Mark out centre lines along each piece, then the dimensions of the open end of the horn, and the appropriate outside dimension of the waveguide, separated by the length of the horn. Join up the ends of these lines and extend them about 2cm past the waveguide end line. See figure 1.

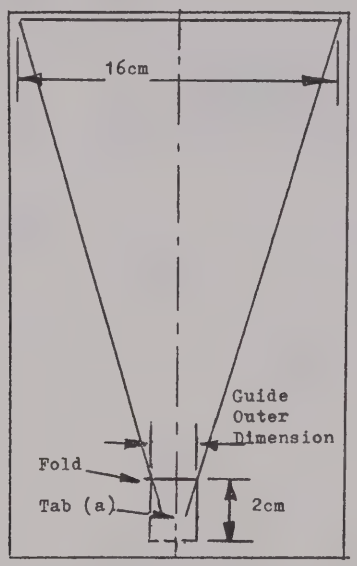


Figure 1

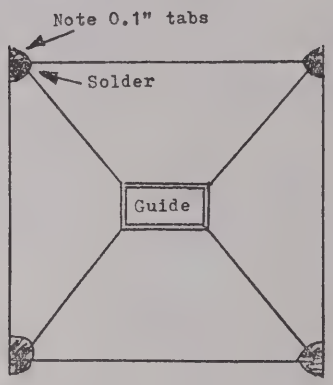


Figure 2

Cut out one pair of opposite sides exactly along the marked lines, on the other pair leave about 0.1 inch extra along the sides. See figure 2. Then fold the tabs (a) back about 10 to 15 degrees. Hold two adjacent sides in position and make several spot solder joins along the seam. Repeat this process untill all the four sides are joined up. Make sure that the narrow end is a snug fit over the waveguide by holding the waveguide in place while doing the spot soldering. Then extend the solder joints all along the four seams, both inside and outside, for rigidity. All this soldering can be done with a medium powered soldering iron.

Then insert the waveguide into the tabs, making sure that they grip it firmly. A gas flame will probably be needed to help solder the waveguide in place. A suggestion is to prop the horn horizontally on a gas stove, with the guide a few inches above the flame. The flame should only be used to keep the metal just below the melting point of the solder, and the soldering iron used to melt the solder locally where required. This avoids the whole structure falling apart when all the solder melts at once! Also, solder the end of the waveguide to the inside of the horn, making a nice smooth join at the throat of the horn. This process is much easier if all the mating surfaces around the guide and tabs are tinned before assembly; if all the surfaces are very clean, and a separate flux is available, this precaution may not be necessary. While the waveguide is hot the flange can also be soldered in place.

When the horn has cooled down, scrape off any remaining flux and give is several coats of paint both inside and outside to prevent it rusting. A red-brown epoxy based paint sold by Halfords for car repairs is recommended for this and other outdoor items as it seems to adhere to surfaces very tenaciously.

USE OF REFLECTORS ON 10 GHz - G3NAQ

The short wavelengths, narrow beam widths, and increasingly sensitive narrowband equipment at 10 GHz makes reflection from objects to span difficult paths a viable proposition. Examples were quoted at the Martlesham Microwave Round Table (19 October 1980) by G3WDG, viz. successful contacts being made via Didcot power station cooling towers and via Fawley oil tanks. Successful tests have also been reported by F1AQS, who also attempts to estimate the field strengths to be expected (DUBUS 4-79 pp.260-2). I hope that the following observations are of interest.

Contrary to what one might guess, the optimum position for a reflector is near one station or the other, and the worst place is midway along the path. This follows firstly from the inverse square law of signal strength; consider the direct path over a distance $D=R_1 + R_2$, where R_1 and R_2 are the distances from each station to the reflector (fig. 2). Then the received power is proportional to:

$$\frac{W.r}{R_1^2 . R_2^2}$$

$$= \frac{W.r}{R_1^2 . D^2 + R_1^4 - 2 R_1^3 . D}$$

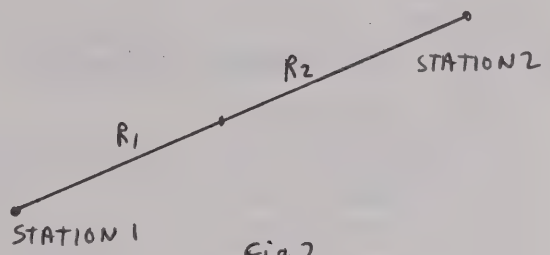


Fig 2

where W is the ERP of the transmitting station and r is the reflection coefficient of the reflector in the appropriate direction. One can show that this function has a minimum at $R_1=R_2$. Obviously if the reflection point is away from the direct path, $R_1 + R_2$ D, and the received signal strength is even weaker (fig. 3).

A second consideration is the "aspect" of the reflector, i.e. the greatest possible area must be visible from both stations. For a cylindrical reflector, assuming that the reflection coefficient does not change with angle, the reflected power falls off as cosine (reflected angle/2), merely from the aspect (fig. 4). Obviously objects irregular in shape will have a much more complicated field pattern, but will also be less likely to give useable signal strengths.

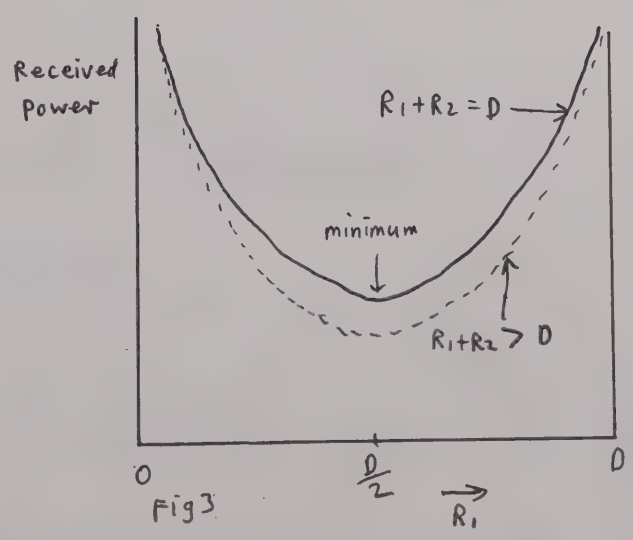
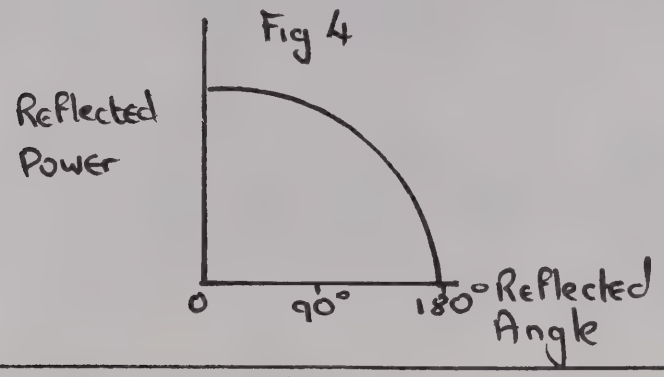


Fig 3

In conclusion, therefore, objects to look for as candidate reflectors should be nearby one of the stations, and preferably away from or in the opposite direction of the direct path, to reflect the strongest signal and present the largest aspect to both stations. Of course, to some extent

these considerations will conflict with the requirement that the reflection spans some intermediate obstacle - and often there will be no choice in the matter! To demonstrate that these considerations do, in fact, hold, I will finally quote two examples; first, the flyswatter antenna is just such a nearby reflector which involves no loss (if it is big enough); secondly, it is common experience that the ghost images on TV sets arise from local buildings, not ones many miles away.

PS. Remember polarization can be modified on reflection - it may be worth rotating one of the antennas to optimise signal strengths.



5.6.2 Propagation Modes

Microwave Propagation

We received a rather interesting letter from Nicolas Hamilton, G8TXG, which is reproduced verbatim below:

"I was interested to see the comment in the October Microwave Newsletter that SHF openings rarely coincide with good V/UHF conditions. I believe this to be because the mechanism of the two is different.

"Good V/UHF 'lifts' are the result of tropospheric reflection from a thin layer of high refractive index deviation at an altitude of 1 to 2km. Unfortunately (from an amateur point of view!) as it is a reflection process, it exhibits a fairly sharp cutoff at the HF end (~3GHz), although it may have some effect on 3cm troposcatter.

"Good SHF 'lifts' are the result of low-level ducting; a typical duct (see A. Picquenard: Radio Wave Propagation, p26) exhibits a gradual cutoff at the LF end around 2GHz. However, below this frequency, the value of k should still be little depressed; these conditions are probabl dismissed by the VHF operator as being 'a bi up'.

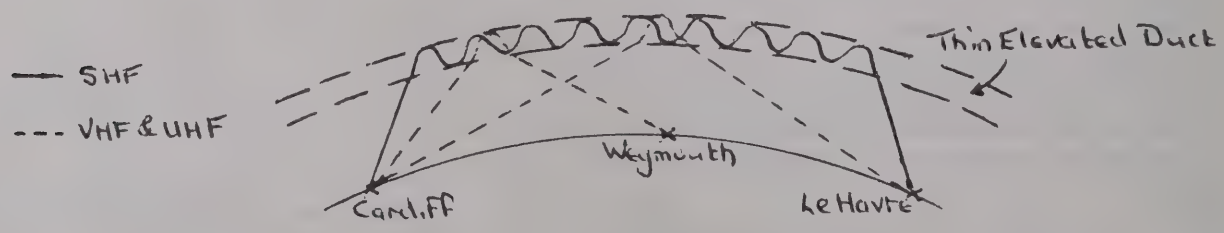
"I conclude, therefore, that the bes indicator for good SHF propagation is not th strength of DX beacons on 144MHz, but monitorin a 70cm beacon/repeater at about 50km range fo abnormal signal strengths. This will indicat abnormal low-level conditions but will b relatively unaffected by tropo reflection.

"If there is a local 10GHz beacon, it migh also be an interesting experiment to use a interferometer system with, say, two verticall stacked horns spaced by a metre or so t identify times when the signal arrives from a abnormal angle."

This is a very interesting theory, and clearl requires some practical investigation. We woul be interested in peoples' comments and (i particular) observations on this.

In the opening on Jan 31st, GW3PPF, located at Cardiff, was unable to exchange 10 GHz signals with stations in Dorset, while able to get into France. However on lower bands (up to 1.3 GHz), he was able to work both areas.

If this was due to the critical angle being exceeded at the top of the inversion layer on the shorter paths, this would in fact cause the lower frequecies to escape first. Thus Phil suggests that perhaps the 10 GHz propagation to France could be due to the formation of an elevated duct of small thickness, such that it is too small to propagate but able to reflect VHF and UHF frequencies.



If this were the true propagation model, this would mean that the best paths to try during openings would be those coinciding with the maximum range obtainable on VHF and UHF, rather than shorter ones.

5.7 Talkback

Talkback

This subject is frequently raised but after the first two cumulatives we ask readers to consider if it in fact does fulfill its purpose: ie

1. Provide information about which stations are out.
2. Enable readers to make contact with any of the stations when they want to try a microwave contact.
3. Provide a reliable "telephone" type engineering link for use while establishing a microwave contact.

When answering these questions it should be bourne in mind that:

1. Whilst high gain beams increase the maximum range in one direction they also reduce the chance of being heard by other stations looking for you.
2. At higher frequencies, higher gain and hence narrower beamwidth antennae are necessary to obtain the same range.
3. Most stations spend very little time actually listening on a calling channel, or even on their talk-back frequency when they are actually trying a microwave contact.
4. As the performance of the microwave equipment has improved, this has demanded much higher performance from the talkback equipment.
5. Higher activity levels and a limited number of "good" sites mean that there are often several stations at one site. This can result in bad qrm on the talkback if they all operate independently, or a very slow qso rate if they all share the same talkback and operate sequentially.
6. Many stations are only out for part of the day and can be missed because a long test or succession of skeds have prevented monitoring of the calling channel for an hour or two.
7. It is certainly possible to go out and never hear stations which could have been worked on the microwave bands.
8. It is not desirable that more effort should be devoted to taking out the talkback system than the microwave equipment!

While 144MHz ssb seems to have become a well established standard other frequencies and modes are occaisionally suggested and used eg 144MHz FM, 432MHz FM or ssb.

We do not think there is an easy solution but comments and ideas are always welcome.

GW3PPF, Phil says 2m is already overcrowded and that talkback is difficult when there are good conditions on the band. He doesn't think moving to 70cm is the answer, because the same problems will only arise again in 2 or 3 years as activity on this band grows. Thus he suggests 23cm as a band where there will be ample space for quite a few years.

Phil says "no doubt S Coast stations will say that during the cumulatives when having cross channel contacts, 2m is quite suitable for talkback. What they may not appreciate is that because they are operating on hills usually above 500ft, to French stations at similar heights, they are not far off line of site to each other. If they all operated down at sea level, they would often find that 10GHz worked even better during good super-refraction conditions, but that 2m would not work at all"! GW3PPF continues by saying that he does not advocate dropping 2m altogether - many people will not accept this. Instead he suggests that initially 23cm used for super-refraction and ducting tests.

5.8. Equipment Performance on the intermediate bands - G3YGF/G4KNZ

Equipment Performance on the Intermediate Bands

Troposcatter path losses are proportional to the cube of the frequency and so a first it might be thought easier to operate on the lower bands, 3.4 and 5.7GHz than on 10GHz. Certainly down on 2m and 70cm people regularly work hundreds of kilometres on a flat band, which supports this assupmtion.

Consider a state-of-the-art narrowband set up on 10 GHz:- say 10W twt to a 4' dish, with 5dB noise figure. Then when operating on 3.4GHz or 5.7GHz the same dish will probably be used (it is not easy to take out anything larger when operating portable). The gain of a given size of dish is proportional to the square of the frequency. So that a 4' dish with 39dBi of gain on 10GHz will have a gain of only 29dBi on 3.4GHz. But the station at the other end of the path will also use a similar size dish, so the net result is that the dish gains drop by a factor of frequency to the fourth power - ie by 20dB when changing from 10GHz to 3.4GHz, for example.

The decreased dish gains is only partially offset by the decreased path loss, which is proportional to frequency cubed - 15dB when going from 10GHz to 3.4 GHz. The net result is that signals will be 5dB worse on 3.4GHz, compared to 10GHz, when running the same power and noise figures on both bands.

Recent tests by GU4KNZ/P and GU3YGF/P with G4MBS on 3.4GHz and 5.7GHz were rather disapointing in comparison with 10GHz, but a second factor is that equipment design is much less advanced at the present (in amateur circles) for these bands. With a power of only 1W on these bands, and noise figures of 15 dB, another 20dB is taken off the received signal level, compared to 10GHz. Thus on 3.4GHz, signals are expected to be weaker by a total of 25dB and when signals are only 15 to 20dBN on 10GHz, this puts them 5 to 10dB below noise. Thus it is not too suprising the path fails.

Others have had similar experiences on 5.7GHz. The calculations thus show that if the same size antenna is to be used on these bands then more attention to equipment performance is needed for the same results (enhanced conditions excepting, of course).

To take this argument one step further, compare a simple 2m talkback station with the 10GHz narrowband station. Here 10W with a 5dB noise figure is easily achieved. But if a small yagi is used (eg 4 ele) with a capture area of say a 4' dish then on a troposcatter path signals will be weaker by the ratio 10/0.144 ie by about 19dB. The talkback is thus demonstrated to be grossly inadequate and this is what is being found by stations listening to 10GHz narrowband signals over long paths. It must be pointed out, however, that the talkback can cover a large area of the country at any given instant, while the 10GHz is only covering a 2 degree sector because of the narrower beamwidth - we are not suggesting working direct on 10GHz!

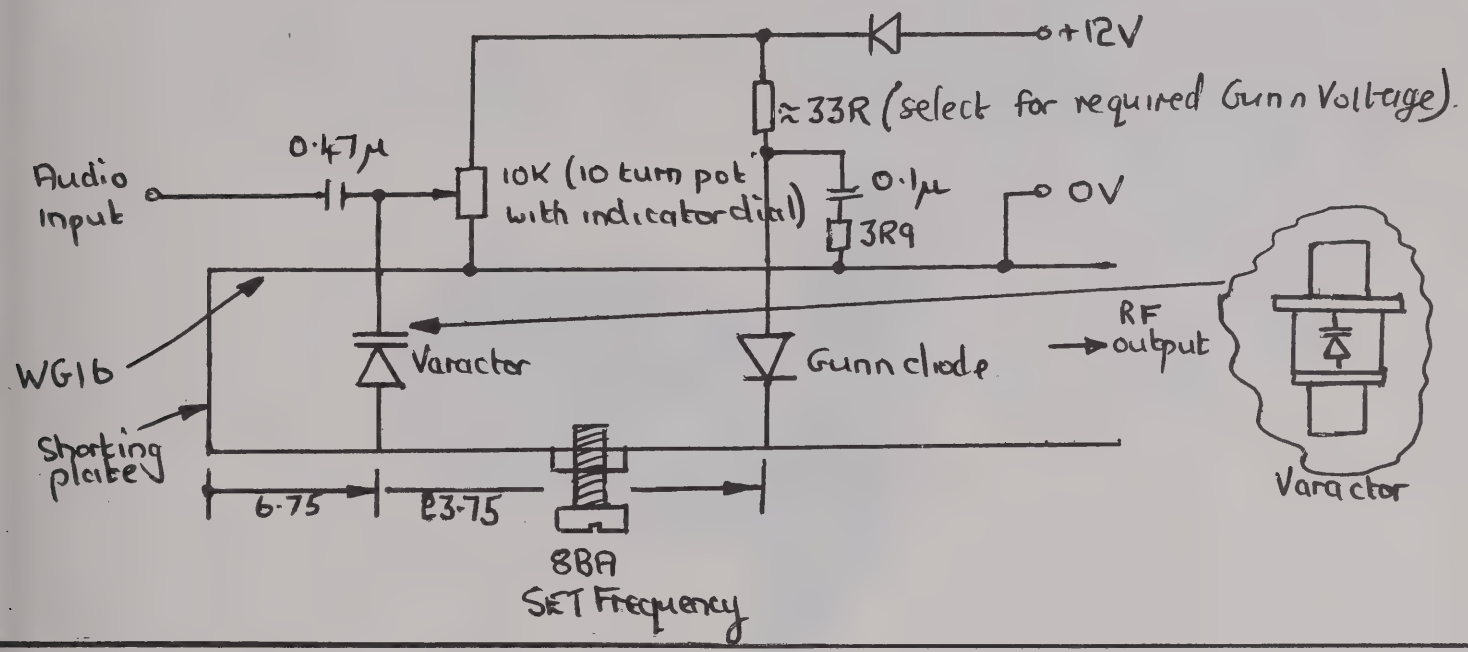
With these facts in mind, G4MBS suggests that it is thus better to try 10GHz first, then try 5.7GHz and 3.4 GHz (assuming similar powers on each band). This is the opposite way to that first thought of. Also the comment "signals are weak on 2m, so 10GHz is probably not worth trying" may result in a lost contact.

Section 6 10GHz Wideband

6.1.1. Varactor Tuning

Varactor Tuning of Gunn Diodes

G4EBF and G8EXL suggest mounting a varactor tuning diode in a cavity behind a Gunn diode for tuning purposes (see diagram). A 12V change in the voltage across the varactor gives a change in frequency of in excess of 100MHz. The diode will typically withstand a reverse bias of 30V, when a change in frequency of about 200MHz should be seen. The dimensions are only suggestions - some experimentation may be necessary. Also, it may be preferred to use a conventional psu to drive and modulate the Gunn, and tune the varactor separately. The exact value of the 33R resistor must be chosen for the individual Gunn diode used.



6.1.2 Mechanical Tuning

G3PFR has tried a nylon screw instead of a PTFE one to tune a Gunn, and has found that it reduces the power output by 10dB and reduces the tuning range. This suggests that PTFE screws may well be useful for tuning the narrowband filters in view of the reasonable performance obtained from nylon screws quoted last month. Does anyone know of a source of them? Mike built two multiplier mounts and filters in a couple of evenings with only hand tools available - he hopes that his success will encourage others to have a go. We could do with a good crop of narrowband beacons!

6.1.3 Mounting

Mounting Diodes

G4DDK suggests a useful item which can be used as a post to mount Gunn, mixer or varactor diodes in waveguide without needing any machining. It is the brass bolt from the old type of ceramic compression mica trimmer. The mounting bolt is removed by breaking the ceramic body; it has a thin disc at the top, suitable for decoupling, and is tapped approx 8 BA for a bolt to pull it to the top of the waveguide; this also serves as a hole for the diode to fit in.

6.1.4 Improvements to self oscillating Mixers

Improved Gunn-mixer receivers

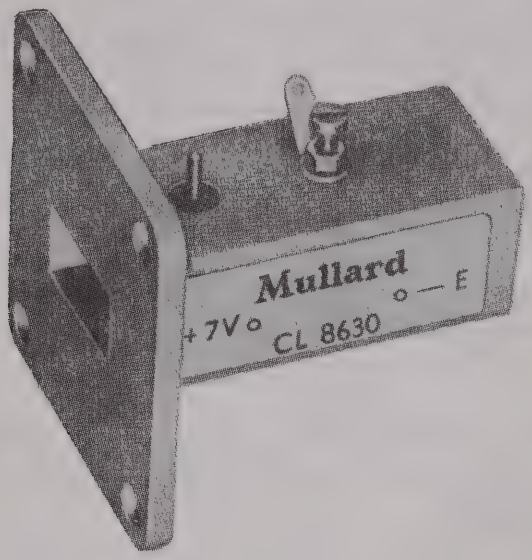
The self-oscillating mixer technique, by reputation a very noisy form of mixer, can be significantly improved by the use of afc. This works by reducing the amount of "jitter", thereby reducing the noise bandwidth and, in turn, the effective noise factor of the mixer. A recent reference to this is *Microwave Journal*, March 1977, pp43-47.

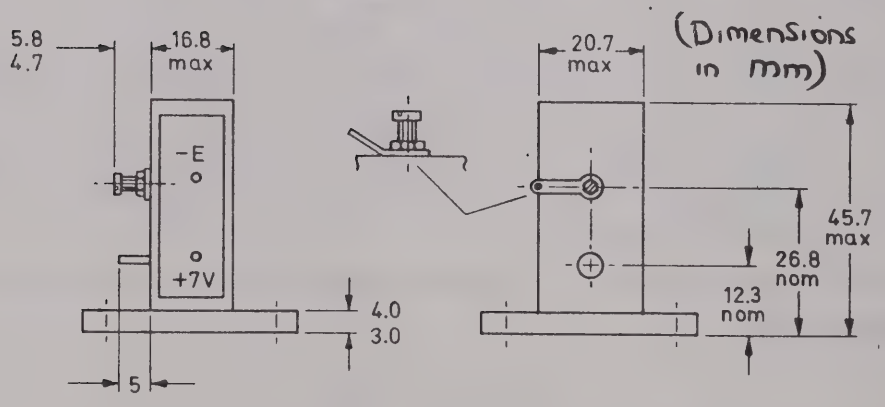
extracted from RadCom

6.1.5

X-Band Gunn Oscillator

At some recent rallies we have seen some Mullard CL8630 Gunn Oscillators on sale. At first glance the waveguide appears too small for a 10GHz Gunn, but the flange mates with a square WG16 flange and it is intentional that there is a step in waveguide size at this point. They come set on a nominal frequency of 10.687GHz, but the tuning screw can be adjusted to pull the frequency down into the amateur band. Usually it is possible to pull them down to 10.1GHz, but certainly there will be no trouble in using them at 10.368 or 10.400GHz. Some data is included below. For anyone interested in designing a Gunn mount, some lessons can be learn from this module as the frequency is remarkably insensitive to the close proximity of objects (eg hands) near the outside, unlike some other designs we have seen.





RATINGS (ABSOLUTE MAXIMUM SYSTEM) at 25°C

Supply voltage max.	+7.5	V
Supply current max. running	160	mA
starting	200	mA
Load v.s.w.r. max.	1.5: 1	

CHARACTERISTICS at 25°C

Centre frequency		10.687		GHz
	Min.	Typ.	Max.	
Power output (at 7.0V)	5.0	8.0	-	mW
Frequency (fixed)	10.675	10.687	10.7	GHz
Frequency temperature coefficient	-	-0.25	-0.4	MHz/°C
Frequency pushing	-	1.5	-	MHz/V
A.M. noise to carrier ratio (1Hz to 100Hz bandwidth)		-94		dB
Second harmonic		-35		dbm

6-2. A Gunn Diode and Oscillator Test Rig - G6XM

The test rig circuit was first devised by G3LQC and
en modified to suit the Solartron CD 1400 Oscilloscope
G6XM. It can be used for the following tests:
Ascertaining the frequency at which the Gunn
cillator is operating.
Selecting the best Gunn diode for maximum output.
Checking that there is no spurious output and that
e diode is operating correctly.
Finding the optimum operating voltage.
There is nothing critical in its construction except
at the leads to the "Y1" and "Y2" plates of the scope
ist be screened to avoid external pick up. The power
upply should be well smoothed also.

Refer to figure 1 and then proceed through the following
eps:
Connect a 1 to 2 Amp 12 Volt power supply to the
st rig and set to about 10 Volt. This level is not
itical, but should not be less than 9 Volt.
. Connect the "X" sweep voltage from the scope as shown.
. Switch the scope on and set "Y1" sensitivity to 2V/cm
d "Y2" sensitivity to 0.2V/cm (This may be altered to
it the scope in use).
. Set the time base of the scope to about 20Hz. Observe
at the lamp LP1 flickers and that a "saw tooth"
veform sweeping from zero to about 10V is shown on
e "Y1" scope trace.
. Ensure the polarity of the Gunn diode is correct
fore connecting.
. Connect the Gunn Oscillator and see if it is oscillating.
trace similar to that shown in figure 2 should be
tained after adjusting the waveguide attenuator. The
timum voltage for best output from the Gunn oscillator
an be seen by inspection of the two scope traces.
. The output from the Gunn oscillator should be smooth
s shown. Any spurious output will be seen immediately
n the "Y2" trace.
. The wavemeter can now be tuned to ascertain the
requency at which the Gunn oscillator is operating.
educe the wavemeter coupling to the minimum required
o show a positive "suckout" point similar to that shown
n figure 2.
. The scope time base speed may be varied but it has
een found that the best speed is below 50Hz.

Figure 2

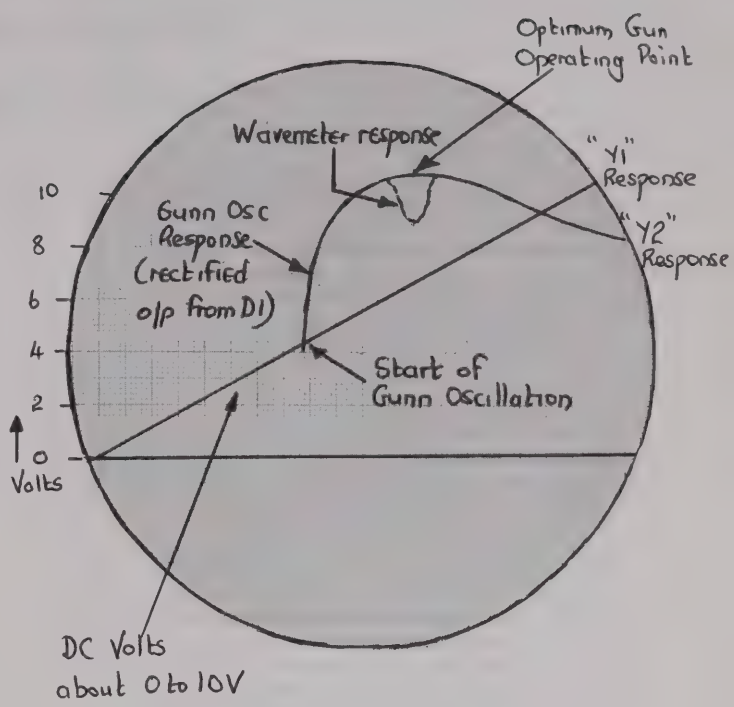
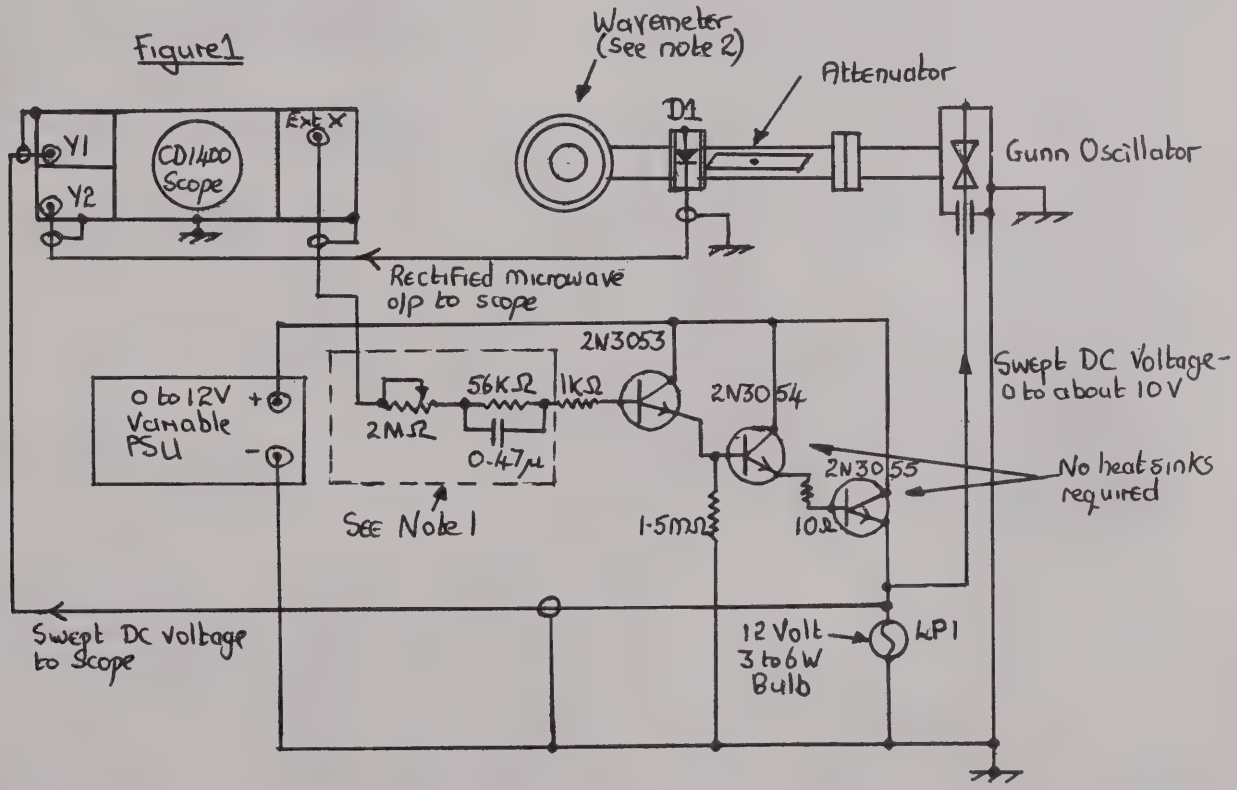


Figure 1



otes
. The X sweep output of the Solartron CD 1400 scope is
igh impedance and the values shown within the dotted
ine gave the best results. For a different scope these
ay have to be changed.
. The wavemeter shown terminates the end of the waveguide.
n "in line" wavemeter should be connected between the
ttenuator and the detector diode D1.

Other scopes with different "X" sweep outputs may need an inverter or gain stage, using an op amp, to provide a suitable sweep signal.

Both G3LQC and G6XM report finding the test rig to be invaluable for finding the best operating conditions for the various Gunn diodes used in their 3cm gear. Many tnx to Bill, G6XM for the info.

6.3 page 1

A simple 10GHz receiver with transmitter option

by D. EVANS, G3RPE, and C. SUCKLING, G3WDG*

Introduction

During recent years amateurs have had much success with simple 10GHz equipment. Usually this has consisted of transmitters generating a mere 1–20mW, receivers using point-contact diode mixers and with an i.f. bandwidth of 100–200kHz, and horn or dish antennas having gains of 20–35dB. Provided that the equipment is operated to take advantage of low-loss propagation modes, such as that over line-of-sight paths or of super-refraction via ducts, then contacts over paths hundreds of kilometres long can be achieved virtually as a matter of routine.

Both calculations and direct measurements show that this size of equipment usually has a reserve of system gain of tens of decibels. This reserve means that it is not even necessary for the equipment to be working well for it to be successful: an overall efficiency of one per cent may be all that is required to cope with most low-loss paths. It also means that "compromise" techniques, such as the use of a Gunn oscillator as a self-oscillating mixer (and usually also as the transmitter), are quite satisfactory under these propagation conditions *provided* that the rest of the system is working reasonably well.

However, for paths containing obstructions the path losses are normally very much greater. Losses 60–80dB (1–100 million times) greater than those over unobstructed paths are not untypical. In order to work over these obstructed paths, the overall system gain has to be correspondingly increased and it becomes necessary to start counting every decibel. The receiver described below, although it is simple to construct, is intended to be efficient at this level. An unusual feature is the flexibility of the design. Because of its particular configuration, a small amount of the local oscillator power is radiated and obviously this can be modulated and used as a transmitter. By a simple modification the output power can be increased if desired, but at some expense of the performance of the receiver.

Design and construction

The receiver is shown schematically in Fig 1. It consists of a simple mixer assembly which is connected directly to a Gunn oscillator of the type which defines its cavity by an iris. The mixer uses a length of waveguide into which is fitted the mixer diode, the hot end of which is decoupled and feeds the i.f. amplifier in the conventional way. Diodes of the 1N23 type are recommended; those with later prefixes (E, F, G) are preferred for their lower noise figures. The signal-input end of the guide can be of any convenient length, and it is fitted with a matching screw or screws to match the mixer diode to the waveguide. The length of the waveguide at the local oscillator end is critical: it needs to be made electrically an odd number of quarter guide wavelengths, ie $n\lambda_g/4$, where n is 1, 3, 5, 7 etc as is convenient. This rear cavity is closed by the same iris as is used to define the Gunn oscillator cavity.

A basic problem in the design of receivers is how to couple the local oscillator drive into the mixer while keeping to a minimum the amount of signal lost by its coupling with the local oscillator circuitry. A feature of the present design is that this isolation is provided simply by using the iris to undercouple the Gunn oscillator. Apart from simplifying construction compared with other methods of achieving isolation, for example by using a directional coupler, the present method has the important advantage of enabling the loaded Q of the Gunn oscillator to be significantly increased. This means that the stability of the Gunn oscillator is improved, which in turn raises the overall efficiency of the receiver.

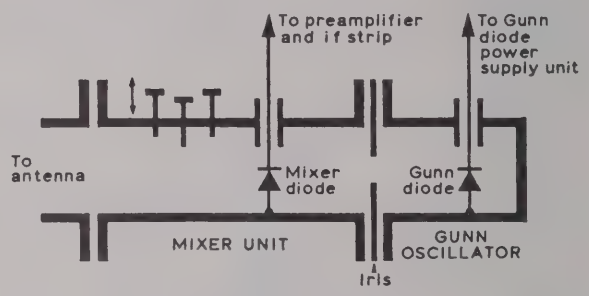


Fig 1. The general arrangement of the receiver

extracted from RadCom

6.3 page 2

Practical details

Two forms of the mixer assembly (which were developed quite independently) are shown in Figs 2(a) and 2(b). Also shown as Fig 2(c) is a recommended design of Gunn oscillator which is the G8APP design [1] with a fixed rather than adjustable rf short. A feature of the design given in Fig 2(a) is that it requires the minimum amount of tools in its fabrication. Points that can be made with respect to its construction are:

(a) First drill a hole about 3/32in diameter centrally through the broad faces of a suitable length of waveguide 16 and open one of the holes to 0·25in diameter.

(b) Remove the brass centre boss from a knob intended to be used with a 0·25in diameter shaft by breaking away the surrounding bakelite. Fit the two flanges in their positions and solder these and the boss in a single operation. The latter may be jigged using a 0·25in drill. Note that the position of the input flange is not critical in any way, but that at the oscillator end it should be within about 1mm of that specified.

(c) Drill and tap the holes for the matching screw. Remove the excess waveguide projecting from the flanges by sawing, filing and finally by grinding on wet silicon carbide paper backed by a sheet of glass. Carefully remove burrs from the inside of the guide, especially where the insulation is to be fitted.

(d) Carefully file away the lip from the mixer diode large connection (or from the adaptor if the diode is of the reversible type) so that the connection is uniformly 0·25in diameter.

(e) Drill the hole in the capacitor plate so that it is a tight fit on the diode pin. When assembling, press the diode against the wall of the guide before tightening the grub screw.

The construction of the design given in Fig 2(b) is similar, but in this case the diode is bolted to the by-pass capacitor at one end, while the other end is made a tight fit in the wall of the guide. In mixer diodes that are reversible it will be found that one connection pin is solid and, preferably, this is the one that is tapped. The pin is undersized for the 8BA thread specified, so the forces involved in tapping the thread are small: it can be done while holding the diode with the fingers.

The fabrication of the Gunn oscillator should present few problems. Constructional details are given in [1] if these are required. The heatsinking in this design is certainly adequate for the low power diodes which generate up to 20mW and which dissipate about 1W. It is insufficient for high power devices which dissipate about 10W. Note that for many low-power Gunn diodes the connection with the flange should be made negative.

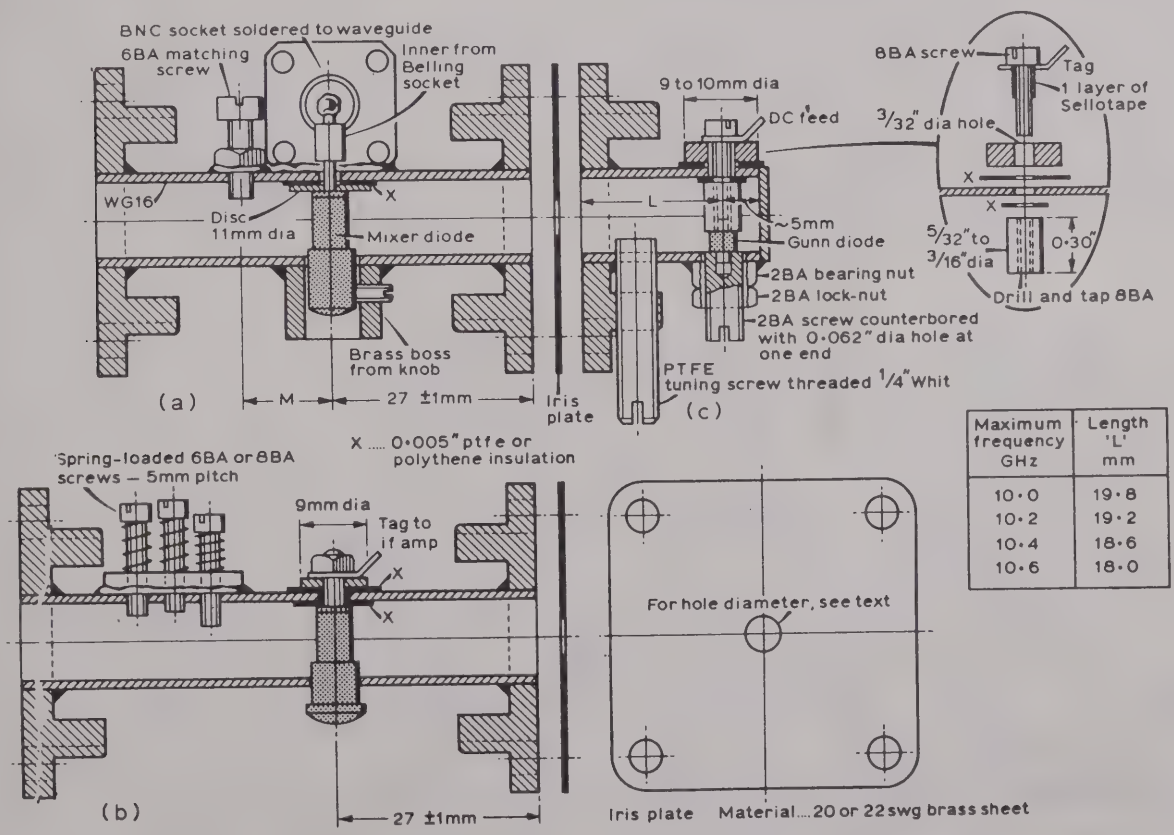
Alignment

The preferred method of aligning the mixer is as follows:

(a) Connect the input of the mixer via a variable attenuator [2] to a suitable rf source, which can conveniently be the local oscillator to be used. Inject rf at *signal* frequency and adjust the matching screws to maximize the mixer current while, at the same time, setting the variable attenuator so that this maximum occurs at the optimum value for the particular mixer diode being used. For point contact diodes, a current of 250–500μA is suitable. The matching screws should then be locked in position. During this operation, the rear end of the mixer cavity should be closed with either the iris to be used or by a blank plate.

(b) With the input connected to a matched load [3], and the Gunn oscillator fitted in its normal position, alter the size of the hole in the iris plate until the diode current is the same as that during *(a)*. Obviously the size of the hole will depend on the output power of the oscillator but will normally be in the range 3–5mm diameter.

In an alternative method used by G3WDG, the receiver is assembled with the antenna and waveguide run which is to be used. For initial tests an iris about 4mm (3/16in) is suggested. The matching screw (in the appropriate hole in the G3WDG design) is then adjusted to set the mixer current at about 250μA. If the current is greater than this, even when the screw does not penetrate into the guide, then the iris should be reduced in diameter. Conversely, if the mixer current obtained with up to the maximum recommended penetration of 3–4mm is still less than the optimum value, then the size of the iris should be increased. If the size exceeds 6mm, then there is a risk that the stability of the Gunn oscillator might be adversely affected. If the mixer current is still too low, then a fault in construction, a poor mixer diode or a badly-matched antenna should be suspected. The latter can be checked by substituting a large horn (or any other well-matched load) for the antenna in question. If correct operation is obtained, then the matching of the original antenna should be improved using, for example, another set of matching screws fitted to the antenna.



Maximum frequency GHz	Length 'L' mm
10·0	19·8
10·2	19·2
10·4	18·6
10·6	18·0

Fig 2. (a) One form of the mixer assembly (G3WDG). The single matching screw fits in one of two holes tapped close to the centre line of the guide, with the length M 11mm for 10·0–10·1GHz and 10mm for 10·3–10·4GHz. (b) An alternative configuration for the mixer (G3RPE). The position of the matching screws is not critical; they can be placed any convenient distance from the diode. (c) A modified G8APP design of Gunn oscillator

Alternative configurations

The critical dimension of the mixer assembly is the length of the guide between the diode and the iris. This was determined experimentally by fabricating an adjustable iris from 0·02in thick sheet 0·9in wide which was bent into the form of a square "U" with the base 0·4in wide. Using the set-up described under Alignment *(a)*, the position of the iris, the penetration of the matching screws and the insertion loss of the attenuator were adjusted at signal frequency to peak at the optimum current for the mixer diode. It was found that moving the iris away from its best position by up to about 1mm could be compensated for by readjustment of the matching screws. The value given in Fig 2, 27mm, represents a compromise length between 10·0 and 10·5GHz. It is somewhat smaller than the values calculated for $3\lambda_g/4$ at these frequencies, namely 29·8 and 27·4 respectively.

The same procedure is recommended if it were desired to optimize the mixer assembly at another frequency, or to lengthen the cavity by making it $5\lambda_g/4$ or $7\lambda_g/4$ in order to fit a wavemeter.

Other Gunn oscillators which employ an iris at the output flange can be substituted directly. Examples are given in [4].

As noted earlier, some of the local oscillator power is radiated from the antenna port and may be used as a low-power transmitter. By increasing the size of the hole in the iris plate the amount radiated may be increased to make the transmitter more effective, although the reduced Q of the oscillator cavity resulting from this change means that the efficiency of the receiver will be impaired. Despite this, the performance of such equipment should be competitive with that of most other transceiver configurations. The size of the iris should not exceed about 6mm diameter, otherwise the stability of the Gunn oscillator may be seriously affected.

Power supply unit

The simplest practical psu consists of a zener diode stabilized circuit as shown in Fig 3. If, as in this case, the receiver local oscillator is not to be modulated, then the working voltage of the Gunn diode will be close to that which produces maximum power output. This can be checked by operating the oscillator via a variable resistor (eg 47Ω 3W) from a 12V 0·3A dc supply, and using the mixer diode current as a power indicator. A zener diode of the optimum working voltage and 1W rating can then be fitted, and the value of resistor R set so that the zener diode passes 50–100mA with the Gunn diode connected.

It is of advantage to be able to frequency modulate the receiver local oscillator with tone since this enables cw signals to be detected. If there is a chance that the unit will be used as a transceiver, if only as a low-power spare equipment, then it is worthwhile also to build in speech modulating facilities. A recently developed circuit produced by G8AGN/G8CZO is given in Fig 4. A deviation of about 100kHz is usually employed, this being suitable for standard broadcast fm components.

Receiver preamplifier and i.f. strip

One of the advantages of having a separate receiver is that there is a wide freedom of choice for the i.f. frequency. In fact almost any frequency can be used in principle since its value will be taken into account during the final calibrating process. With single-ended mixers of the type described above, there are advantages in using a relatively high i.f. in the region of 100MHz. A convenient form for this can be a standard fm broadcast receiver, which may have limited afc facilities built in—as well as a tape recorder. Some receivers of this type can tune 150MHz, which is a useful i.f. in that the receiver can be made to tune 10,000–10,100MHz on one channel, and 10,300–10,400MHz on the other. Suitable preamplifiers using either BFY90 or 40673 devices are given in [5].

A useful check on the overall performance of the receiver is to measure the difference in its noise output when the antenna is pointed at the sky or at objects such as the ground. This technique is described briefly in [6].

References

RC = *Radio Communication*; *M* = RSGB *VHF/UHF Manual*, 3rd edn.

[1] *RC* February 1976, p123.
[2] *RC* December 1972, p280; also *M*, Fig 8.39.
[3] *RC* November 1972, p741; also *M*, Fig 8.36 and 8.37. Horn antennas normally represent a well-matched load.
[4] *RC* May 1974, p288, Figs 6, 7 and 8. Figs 7 and 8 are also shown as Figs 8.52 and 8.53 in *M*. Also *RC* September 1976, p667, Figs 1 and 2.
[5] *M*, Figs 8.74—8.78.
[6] *RC* July 1977, p541.

□

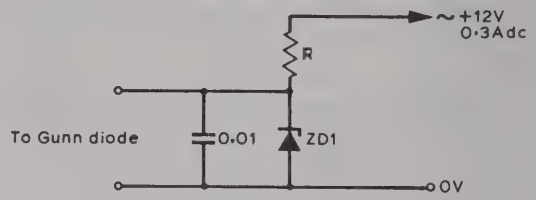


Fig 3. A simple zener diode psu. The working voltage of ZD1 is normally 7–9V, see text

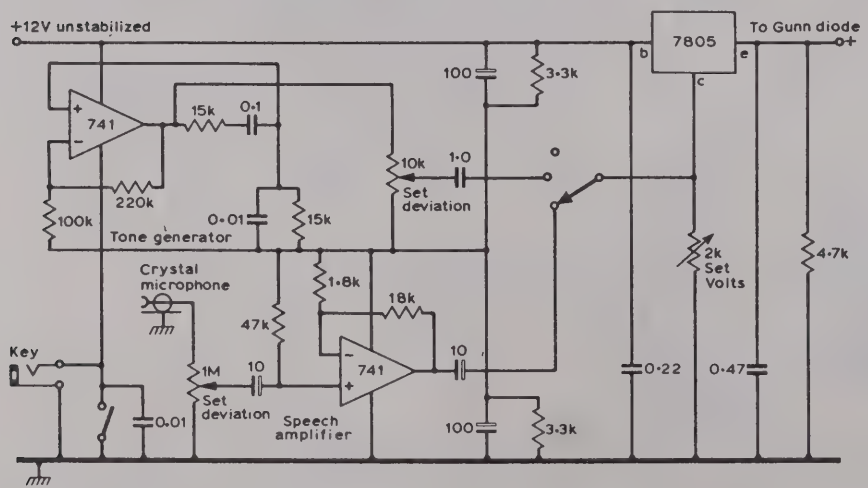


Fig 4. A modulated psu which is due to G8AGN/G8CZO

6.4

A Doppler Transceiver for 10GHz - G3YGF

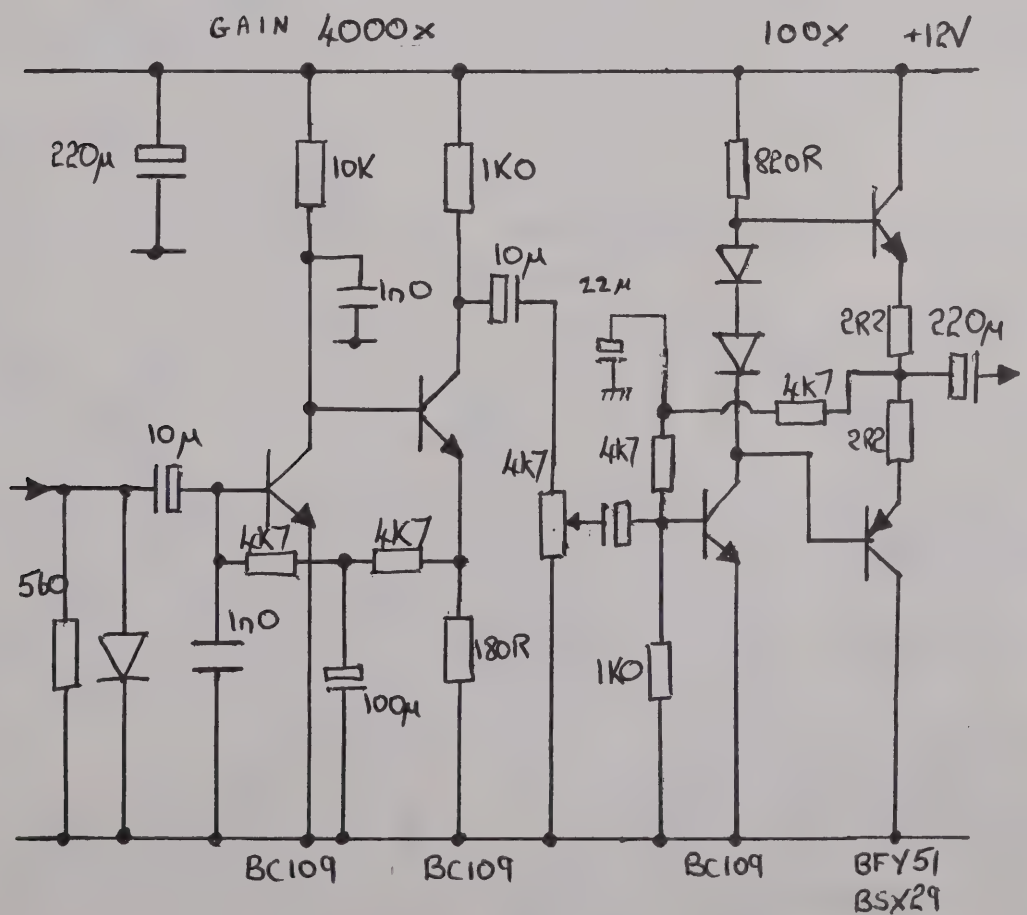
Heavy rain and weather fronts have been shown to enhance signal levels on 10GHz over obstructed paths by acting as scattering objects for the signals. When this happens the rapidly moving particles may spread a narrowband cw/ssb signal out over several kiloHertz because of the doppler effect.

It is desirable to be able to detect this type of weather by using one's own equipment, without requiring a constant signal from another station. G4MBS has been doing this by using a low power narrowband transmitter at his home, and listening for it on a separate receiver using aerials with fairly broad beams. A scattering cloud can be heard either by the increased signal strength from the transmitter, or by listening for the doppler shifted components of the signal.

It occurred to G3YGF that it might be possible to use a much simpler setup for this purpose, such as a Gunn type intruder alarm, after being 'detected' by one of these one day in a shop. It would probably not be quite as sensitive as the narrowband setup, but considerably simpler.

Doppler shifted signals will be seen only when the target is moving towards or away from the antenna, so the antenna should be angled upwards slightly for best results. A target velocity of 30mph will give an audio tone of about 600Hz, so any rain will be heard as signals in the low part of the audio spectrum. With a 20dB horn, it will detect cars at about half a mile range. Besides this use, G3YGF has discovered that quite a lot can be learnt about microwaves and standing wave patterns just by playing with a very simple rig such as this and observing various moving objects: cars, fan blades, birds, trees etc.

If a doppler module is not available, an existing wideband rig can very easily be modified by replacing the IF amplifier on the detector diode by the audio one shown below. The circuit has sufficient gain to drive a loudspeaker directly. The leads from the mixer to the amplifier should be kept short and screened to prevent hum pickup, and the supply may need more decoupling if it is noisy, eg a car battery when the engine is running. The frequency response extends down to low audio frequencies, and it is probably limited by the speaker or headphones in practice. Slightly better results can be obtained by using a 3dB coupler in place of the 10dB one usually used in a receiver, as this gives the best compromise between the amount of oscillator power lost in the dummy load on the coupler and the amount of received signal lost in the oscillator.



6.5 Gunn Diode Power Supply G4KNZ/G3YGF

There have been a number of designs and ideas for Gunn Diode power supplies in Radcom, the VHF/UHF manual and elsewhere. The following design is an attempt to collate these ideas into a simple, reliable and tested circuit requiring a minimum of commonly availa components.

Design:

1 The basis of the power supply is the 7805 regulator IC with a variable resistor between the common terminal and earth to make the output voltage adjustable in the rang 6 - 9 V. It is also protected against output short circuits of short duration, but the heat sink is not capable of withstanding this condition indefinitely. Modulation from a mic amplifier or tone oscillator is applied to the common terminal of the IC via a capacitor. The level required depends on the Gunn voltage, but is typically from 10 to 100 mV peak to peak.

2 The mic amp is a single transistor circuit providing a gain of about 10 times, and is suitable for use with a low impedance mic.

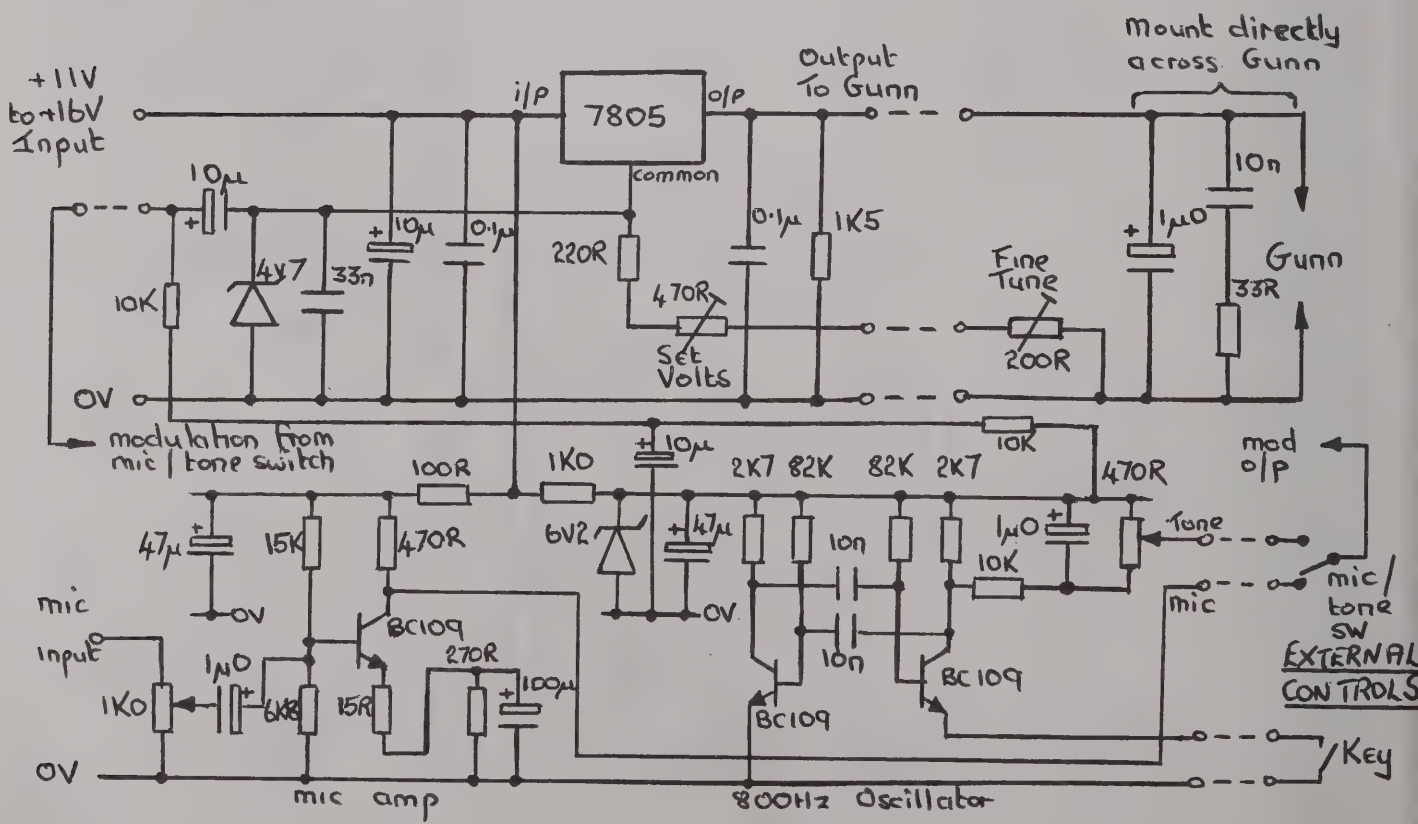
3 The tone generator is a two transistor oscillator which can be keyed on and off for CW.

4 Several decoupling capacitors and resistors in the supplies have been added to prevent noise on the power supply or tone from the oscillator from reaching the mic amp or the Gunn Diode.

5 Every attempt has been made to keep the circuit as simple as possible, and so no op - amps have been used. There are only two types of semiconductors used - the regulat IC and some NPN transistors, BC109 or similar.

The following circuit uses component values which will produce an output between 6 - 9 V. A preset on the PCB sets the approximate voltage, the external pot provides a fine tuning facility on receive and has a range of about 1V. Obviously the ratio of the values of these two pots can be changed to give a larger or smaller tuning range, but the total value should remain the same.

The PCB is double sided, with an earth plane on top, size 90mm x 45mm with mountin holes at each corner. The regulator chip is mounted on the board with a heatsink made from a piece of aluminium. This is ok for low currents up to about 200mA, but for larger currents it should be mounted on a sepatate heatsink, or the side of the box. If anyone wants a copy of the artwork to make their own PCBs, please contact G4KNZ.



A 10.7MHz IF Preamplifier and Discriminator - G4KNZ/G3YGF

A number of designs for WB IF strips have been lished in the past in Radcom, the VHF/UHF manual and ewhere. Frequencies of 10.7MHz, 30MHz and 100MHz are monly used, occaisionally others such as 144MHz, ch leaves a wide choice of IF frequencies.

For the beginner, one simple means of obtaining an strip is to use an FM VHF broadcast band receiver, sibly with an external preamp. However, if the IF cuitry is to be home built then probably the plest choice of IF frequency is 10.7MHz. No freq- y conversion is necessary, bandpass filters (eg amic filters) are easily obtained and IC's are ilable for the discriminator. The circuits here are ended to fill the gap of a simple IF strip for 10.7 using modern components.

Most of the circuitry should be constructed on one rd - the ceramic filter, FM discriminator and an io amplifier and this should be mounted in a diecast with the external controls required. For best formance the preamp should be mounted on or close the waveguide in a separate small box. Note that IF transformers in the preamp form an essential t of the filtering and a broad band type will not suitable as an alternative.

In addition to notes on the design and construction ie data on the IC's used is included.

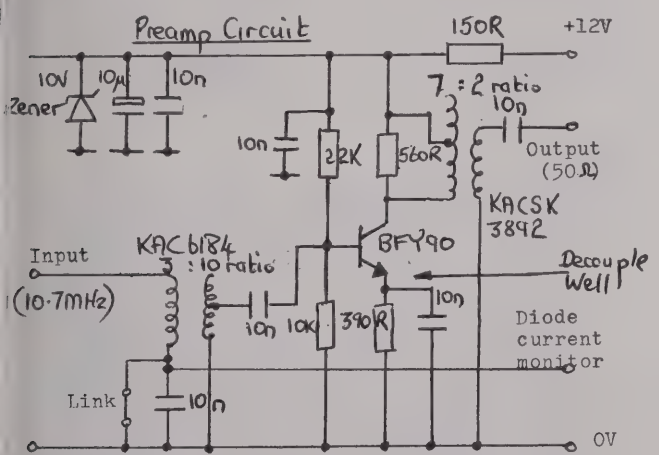
amp Design

A single BFY90 provides a gain of approximately 35

An IF transformer matches the diode impedance of a v hundred ohms to the transistor base of about 1K

The output impedance of the preamp is 50 ohm, an IF ansformer transforming the collector resistor of 560R. Adequate decoupling is essential. In particular, e should be taken to decouple the emitter very close the transistor or the preamp may oscillate. Diode current can be monitored by breaking the link m the input IF transformer to earth and inserting a er.

Supply voltage should be in the range 11.5 to 16V. ier supplies from 10V upwards can be used but change e resistor in the supply to ensure there is 10V on e zener but that not too much current flowing in it.



Strip Design (see following page)

The 300 ohm impedance of the ceramic filter is ched to the 50 ohm input impedance using an IF nsformer.

The output from the filter is fed into an LM3089 CA3089 which provides a complete FM receiver IF stem as one chip.

A single tuned discriminator coil is used across ns 8 and 9 of the 3089.

The audio output is fed via a volume control into LM380 audio power amplifier. The output of up to 5W is suitable for driving an 8 ohm or 16 ohm udspeaker or headphones.

The AFC and signal strength outputs can be used required.

Again adequate decoupling is essential. In particular large electrolytic capacitor is needed to decouple the wer supply for the audio amp.

The 3R3 resistor in series with 0.1uH capacitor from e audio output of the LM380 to earth is necessary prevent the IC oscillating at a few MHz.

The power supply should be in the range 11.5V to V. Note that 16V is the maximum rating of the 3089 or CA3089, so don't use anything too close to e upper value for safety.

System Design

1. The main selectivity is provided by the ceramic filter with a 180kHz bandwidth.
2. The ceramic filter has a spurious response only 25dB down on the main response at about 3.5MHz and so additional selectivity provided by the IFT's is necessary. See plots of the responses.
3. The gain provided by the preamp is sufficient to ensure that the noise figure is virtually that of the preamp ie negligible second stage noise. Too much gain is undesirable, however, as it will make the receiver more susceptible to strong signals off frequency (other stations on site).

Notes

1. All the inductors/IF transformers and the ceramic filter should be available from Ambit International at 200 North Service Road, Brentwood, Essex, CM14 4SG, telephone 0277-230909. The IFT's are around 35 to 40p, and the filter about 50p but it is worth checking on these prices and the delivery time first. The IC's are also in the Ambit catalogue.
2. The circuits should preferably be constructed on a double sided pcb, the earth plane on the top side. With care a "3D" type of construction could be used instead, using the components in mid air above an earth plane. Veroboard and similar materials should not be used.
3. The screening of the circuits is important, otherwise 10MHz broadcast signals are easily picked up. This is a good reason for mounting the preamp close to the mixer diode, on the waveguide.
4. The signal strength output can drive an S meter which is most useful for aligning dishes.
5. A low impedance path for the mixer diode current is important - either a short or low resistance meter. Otherwise a drastically worse performance is obtained.
5. G4KNZ hopes to have the artwork for pcb's for both the preamp and IF strip available if anyone is interested, although this will probably not be for a few weeks.

Electrical Characteristics of the LM3089 & LM380

Electrical Characteristics (T_A = 25°C, V_{CC} = +12V) — LM3089

PARAMETER		CONDITIONS	MIN	TYP	MAX	UNITS
DC CHARACTERISTICS (V_{IN} = 0, NOT MUTED)						
I_{11}	Supply Current		16	23	30	mA
V1, 2, 3	IF Input and Bias		1.2	1.9	2.4	V
V6	Audio Output		5.0	5.6	6.0	V
V7	AFC Output		5.0	5.6	6.0	V
V10	Reference Bias		5.0	5.6	6.0	V
V12	Mute Control		5.0	5.4	6.0	V
V13	IF Level			0	0.5	V
V15	Delayed AGC		4.2	4.7	5.3	V
DYNAMIC CHARACTERISTICS f_o = 10.7 MHz, Δf = ±75 kHz @ 400 Hz						
V_{IN}(LIM)	Input Limiting -3 dB			12	25	µV
AMR	AM Rejection	V_{IN} = 100 mV, AM: 30%	45	55		-dB
V_O(AF)	Recovered Audio	V_{IN} = 10 mV	300	400	500	mVrms
THD	Total Harmonic Distortion					
	Single Tuned (Note 1)	V_{IN} = 100 mV		0.5	1.0	%
	Double Tuned (Note 1)	V_{IN} = 100 mV		0.1	0.3	%
S+N/N	Signal to Noise Ratio	V_{IN} = 100 mV	60	70		dB
V12	Mute Control	V_{IN} = 100 mV		0	0.5	V
V13	IF Level	V_{IN} = 100 mV	4.0	5.0	6.0	V
V13	IF Level	V_{IN} = 500µV	1.0	1.5	2.0	V
V15	Delayed AGC	V_{IN} = 100 mV		0.1	0.5	V
V15	Delayed AGC	V_{IN} = 30 mV		2.5		V
V_O(AF)	Audio Muted	V_{IN} = 100 mV, V5 = +2.5V		60		-dB

Note 1: Distortion is a function of quadrature coil used.

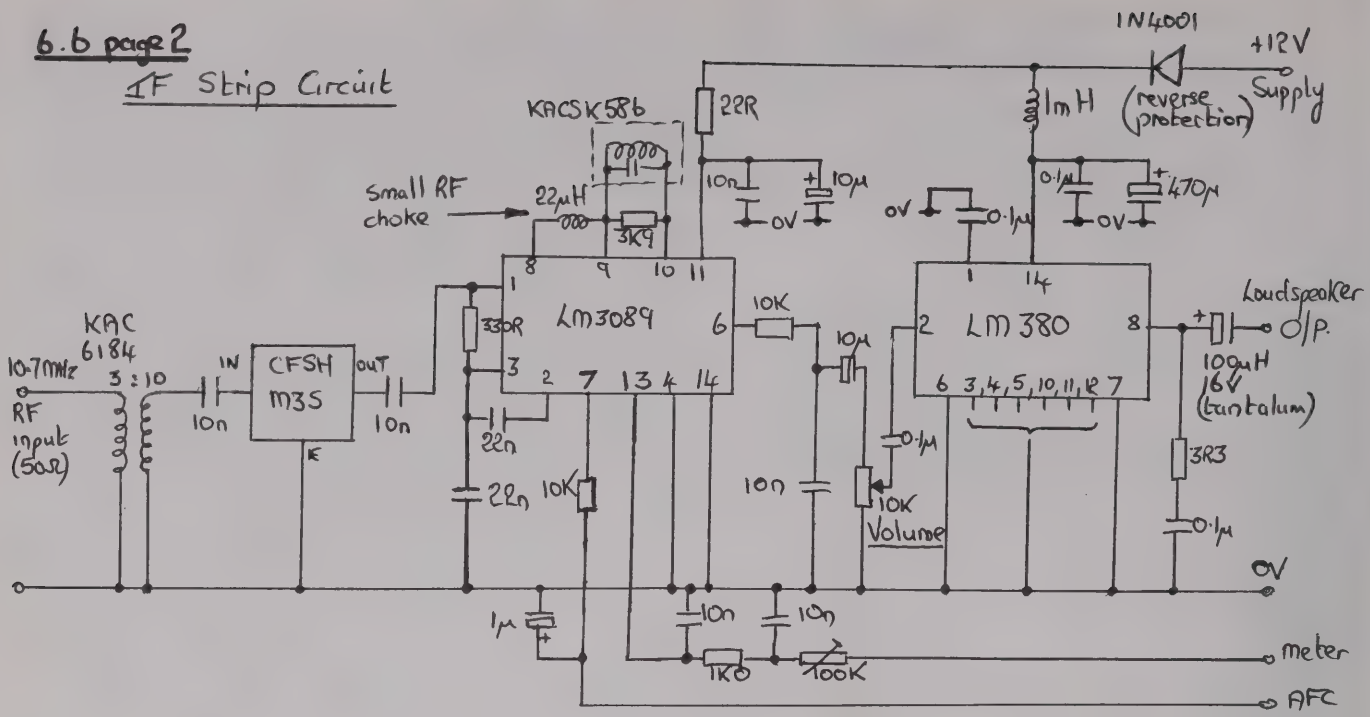
Electrical Characteristics (Note 1) — LM380

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
Output Power	$P_{OUT(RMS)}$	(Notes 3, 4) R_L = 8Ω, THD = 3%	2.5			W
Gain	A_V		40	50	60	V/V
Output Voltage Swing	V_{OUT}	R_L = 8Ω		14		V_{p-p}
Input Resistance	Z_{IN}			150k		Ω
Total Harmonic Distortion	THD	(Note 4, 5)		0.2		%
Power Supply Rejection Ratio	PSRR	(Note 2)		38		dB
Supply Voltage	V_S		8		22	V
Bandwidth	BW	P_{OUT} = 2W, R_L = 8Ω		100k		Hz
Quiescent Supply Current	I_Q			7	25	mA
Quiescent Output Voltage	V_{OUTQ}		8	9.0	10	V
Bias Current	I_{BIAS}	Inputs Floating		100		nA
Short Circuit Current	I_{SC}			1.3		A

Note 1: V_S = 18V and T_A = 25°C unless otherwise specified.
Note 2: Rejection ratio referred to the output with C_{BYPASS} = 5 µF.
Note 3: With device Pins 3, 4, 5, 10, 11, 12 soldered into a 1/16" epoxy glass board with 2 ounce copper foil with a minimum surface of 6 square inches.
Note 4: If oscillation exists under some load conditions, add 2.7Ω and 0.1 µfd series network from Pin 8 to Gnd.
Note 5: C_{BYPASS} = 0.47 µfd on Pin 1.

6.6 page 2

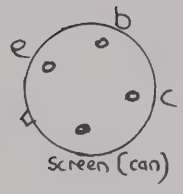
IF Strip Circuit



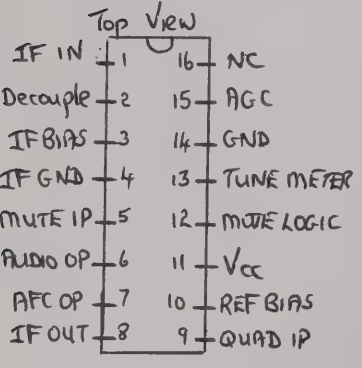
Pin Connections

BFY90

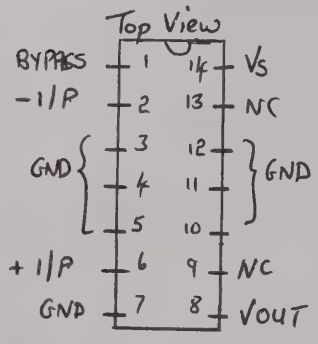
Underside view



Earth can to pcb or earth plane



LM3089N

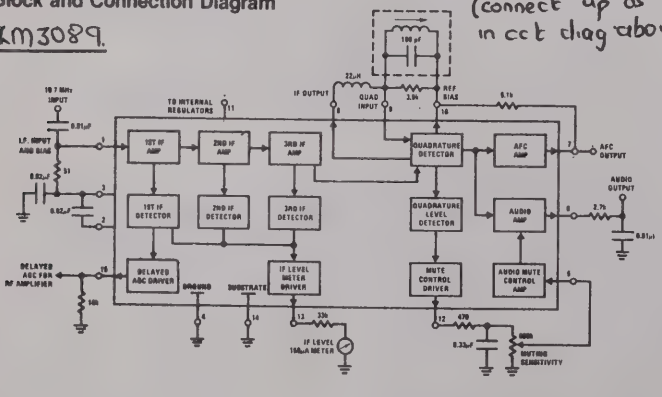


LM380N

Block and Connection Diagram

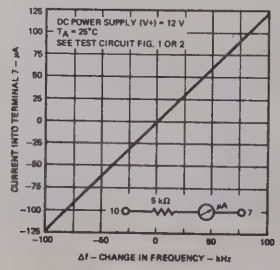
LM3089.

(connect up as in cct diag above)

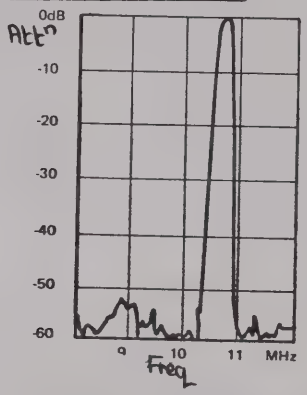


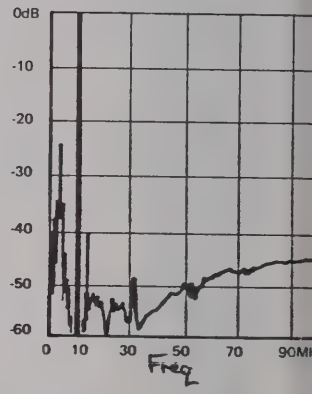
LM3089

AFC CHARACTERISTICS (CURRENT AT TERM. 7 AS A FUNCTION OF CHANGE IN FREQUENCY)

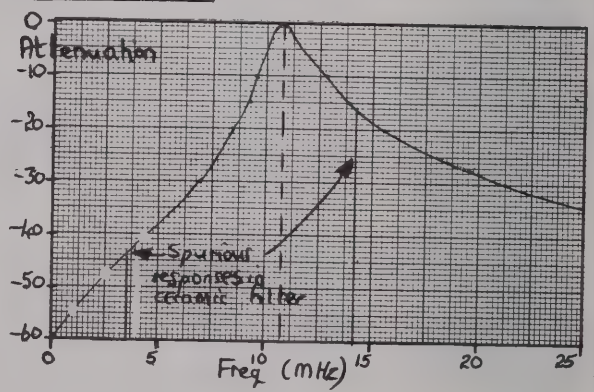


Response of CFSHM3S





Response of Preamp



Some additional rejection of the spurii in the ceramic filter pass band will be provided by the input IFT to the filter.

The performance is not the ultimate that can be achieved but the circuits have been kept as simple as possible.

Possible improvements are the provision of more filtering, with another transistor stage to overcome the additional losses.

The circuits could be adapted for other frequencies but note that the LM3089 will not work much above 20MHz.

Measurements

1. Preamp noise figure measured at 3.7dB but from 50
2. Preamp gain measured at 35dB.
3. Preamp 3dB bandwidth measured at 600kHz
4. Preamp power requirement ~15mA from 12V
5. IF Strip input for 10dB S/N = -83dBm
6. IF Strip power requirement 35mA from 12V, no sig i
7. Overall, for 10dB S/N, required input = -108dBm
8. Overall 3dB bandwidth = 150kHz.

Section 7 10GHz Narrowband

7.1 Varactor Multipliers

7.1.1 Low-power varactor multipliers to 10GHz

Several people have been using redundant Gunn diode oscillator cavities as the basis of varactor multipliers to 10GHz. A typical version is that due to G8DEK. A BXY41B varactor (which is available on the surplus market) replaces the Gunn diode in the cavity shown in Fig 3 on p289 of the May 1974 *Radio Communication*. The simple matching circuit between driver and varactor is shown in Fig 1, and this should be contained within a trough or screening box. Other forms of cavity should work as well provided that the inductance L is optimized.

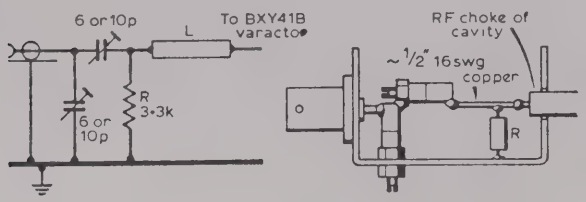


Fig 1. Circuit and layout of driver/varactor matching network

From 400mW of drive at 1·006GHz, an output of about 6mW at 10·060GHz is obtained, which is more than enough to form the local oscillator of a receiver. It would also act as quite a potent transmitter: assuming the use of a narrow-band receiver and operation in an a.m. mode (thus avoiding the 10dB fm penalty), this power is equivalent to at least several hundred milliwatts of a wide-band transmission. It is of course essential that the output of the multiplier be adequately filtered and that the purity of the output be checked, preferably with a spectrum analyser. A practical filter was described in this column in August 1973.

G8APP uses a similar approach, and obtains about one milliwatt from 400mW drive at 400MHz. He recommends that the driver be mounted directly on to the multiplier, as his experience is that movement of any flexible coaxial cable connecting the two can upset the tuning. □

extracted from RadCom

7.1.2 A 10GHz varactor multiplier

A single varactor multiplier driven by a crystal-controlled source at a much lower frequency, a 432MHz transmitter for example, would appear to be an attractive way of getting on 10GHz. However, there are two problems with this type of multiplier system. The first is the vulnerability to changes in operating conditions: minor variations in drive level or in their load, for instance, can cause the output to break up into a large number of spurii. This effect may be difficult to detect, let alone cure, without the use of a spectrum analyser, an instrument to which not many amateurs are likely to have ready access. The second problem is that the output, even though crystal-controlled, may have a (perhaps surprisingly) large bandwidth, and therefore may be far from "narrow-band" by the standards of lower frequencies.

G8DEK has continued his earlier work (briefly covered in July 1975 *Microwaves*) in trying to tame this technique and make it safer for general use. He appears to have made some progress in this direction: when he tested two versions of the multiplier circuit shown in Figs 1 and 2, each with four diodes, it was not found possible to make the output break up under any conditions of normal operation.

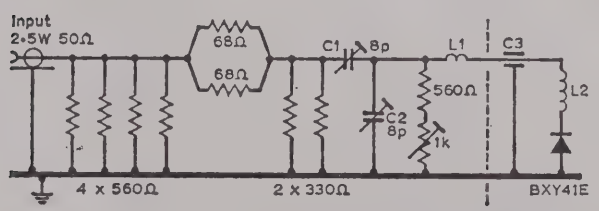


Fig 1. Circuit of multiplier. All resistors ¼W carbon, leads to be kept as short as practical. L1 1¼ t 4mm id; C1, C2 Mullard film dielectric capacitors; C3 capacitor formed between diode post and wall of guide; L2 inductance of diode post

An important feature of the design is the carefully-tuned impulse generator represented in Fig 1 by L2/C3. This has been optimized for the BXY41E snap varactor diode and may not work properly with other diodes. A second feature is the 6dB attenuator which is fitted between the varactor circuit proper and the driver in order to reduce the amount of interaction between the two. It appears that many of the problems with this type of multiplier are due to instability in the preceding driver stage caused by the widely-varying load represented by the varactor. The attenuator of course contributes to the low overall efficiency of the circuit (a characteristic which amateurs as a group appear unduly reluctant to accept), but this seems to be a small price to pay for reliability of operation. For a drive of about 2·5W at 350-550MHz, about 2-4mW of rf at 10GHz can be obtained, which is sufficient for the local oscillator of a receiver, for a small transmitter or as the driver to a travelling-wave tube.

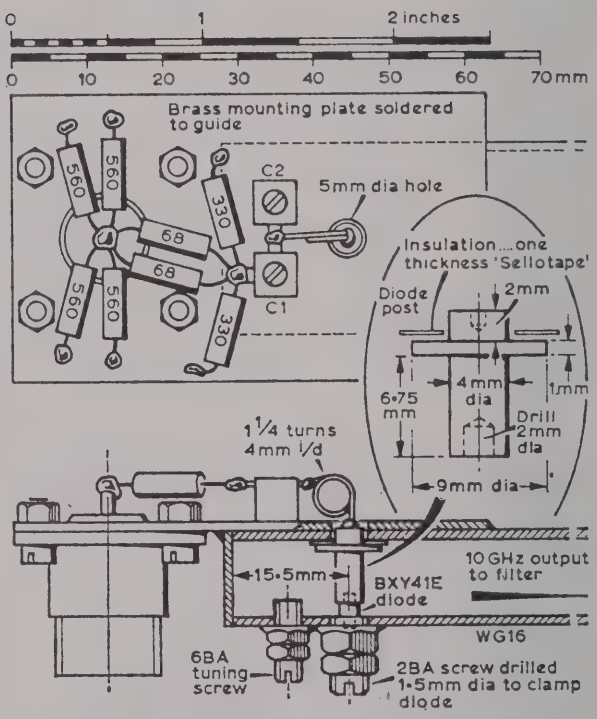


Fig 2. Layout of main components of the multiplier

A filter is essential to remove harmonics from the output: the one described in August 1973 *Microwaves* has been used although its performance is barely adequate in this application.

The second major problem with this approach to getting on 10GHz is the noise bandwidth of the 10GHz output. This noise originates in the oscillator and early multiplier circuits and is always present. However, it becomes much more apparent at 10GHz because the bandwidth is proportional to the square of the overall multiplication factor. Even with the least noisy driver available, which used a 101 MHz neutralized glass crystal in a carefully-designed Butler circuit, the noise measured by G8DEK from a single-ended mixer was about 10dB worse than a Gunn oscillator when using a 10·7MHz i.f. and about 7dB when using a 30MHz i.f. Clearly, there will have to be a big improvement in the quality of drivers before the full advantages of going truly narrow-band are fully realized.

extracted from RadCom

7.1.3 Higher Powers on Narrowband

The step recovery diode multiplier by G8DEK will give around 5mW out at 10GHz when driven with about 1W at 384MHz.

A much higher output power is possible with the same diode, if the multiplier is driven with 1W at 1152MHz, as the multiplication factor is lower.

We do not have any figures for the performance of the DEK design driven at this frequency, but the VHF Comms one with the diode and matching network in a box on a pillar on the guide will give around 50mW out for 1W drive at 1152MHz. This is obviously a simple means of getting higher powers at 10GHz.

The problem is in getting from 384MHz where well proven designs are available, up to similar power levels at 1152MHz where transistors are still a bit touchy and expensive.

While varactor multipliers are always suggested, they tend to be somewhat unpredictable beasts, and can often produce power everywhere except where it it wanted! - Particularly when used over a wide temperature range, and with varying source and load impedances.

The design below, although it uses a varactor, is suggested as a way to higher powers. It can be driven by a BGY22 directly, with about 2W. However it may be more docile if a pad, say 3dB, is used with a higher power source at 384MHz.

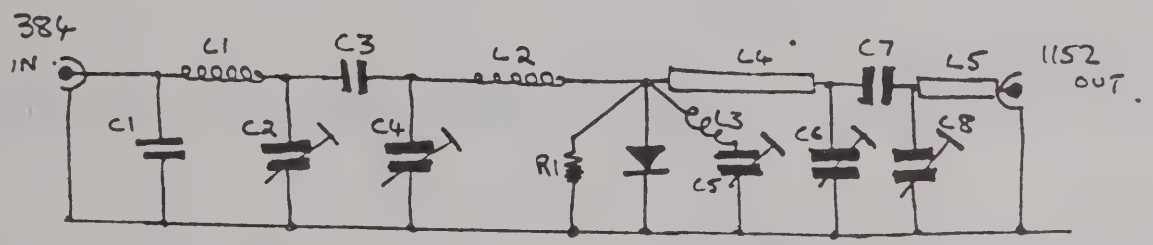
We would very much like to hear about any designs for transistor multipliers which use cheap transistors and give 100mW to 1W out at 1152MHz.

7.1.4 page1 A Varactor tripler from 384MHz to 1152MHz G4MBS

The design is based on one in Nov 74 Radcom. G4MBS has built several of them and found them to be quite tame and reliable.

Performance: 2W input at 384MHz gave 1.2W out at 1152MHz.

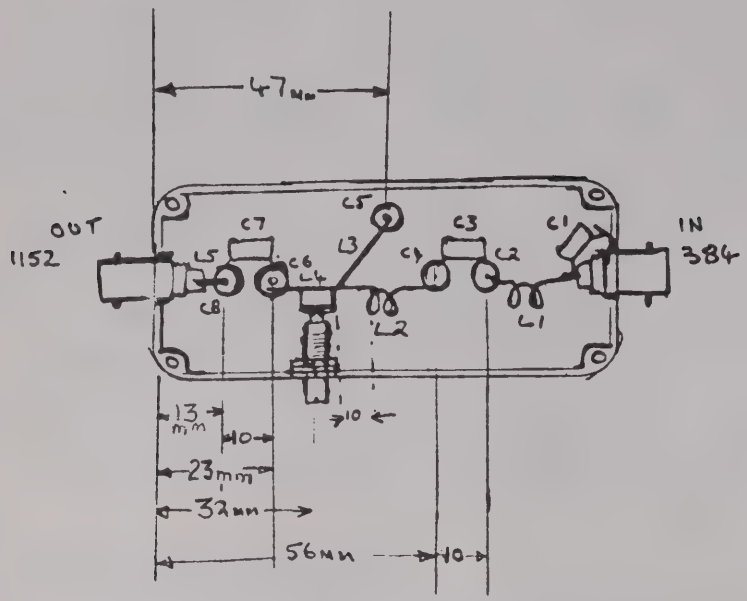
Circuit Diagram:



Parts list:

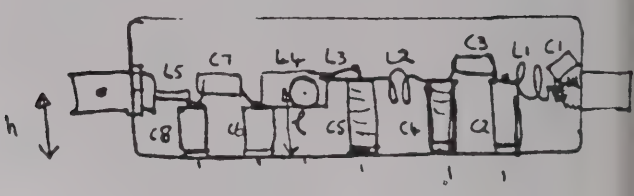
C1	10pF fixed
C2,4,5	10pF tubular trimmer
C3	3pF fixed
C6,8	5pF tubular trimmer
C7	0.5pF fixed
L1,2	2 turns 18swg on 1/4" dia. Coil 10mm long between start and finish of turns, with 3mm leads at each end. The varactor end of L2 has a 10mm lead.
L3	15mm of straight 18swg wire.
L4	Brass strip 16mm x 8mm
L5	Uncut centre connection of a BNC socket!
Diode	BXY27
R1	Pencil lead rubbed on diode body to give about 10k. (perhaps try a preset pot? its a bit less barbaric, and easier to optimise?! - G3YGF)
Box	Eddystone 7969P approx 38 x 93 x 30 mm

Construction:



Top View

SCALE APPROX
0·7 of Full size



Side View

The heights of the sockets (h) and the diode (l) are determined by the height of the trimmers C6 and C8.

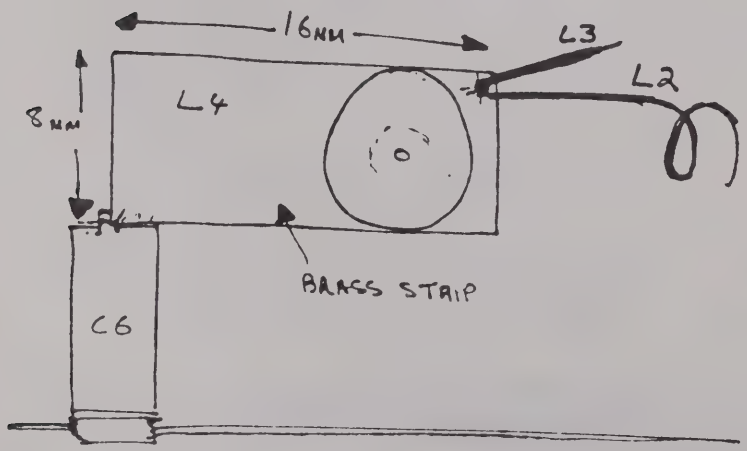
The diode is held in position by a OBA screw which runs in a thread tapped in the side of the diecast box. A locknut is used on the inside of the box.

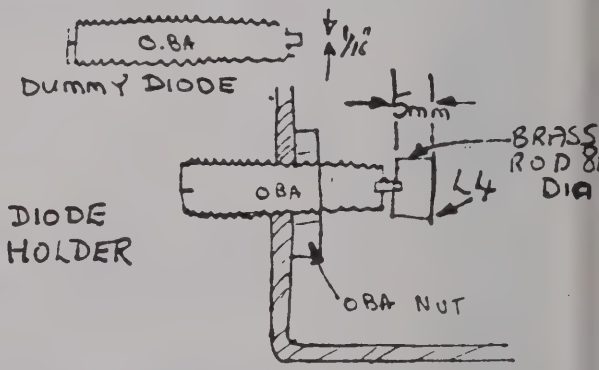
NB: The cylindrical block soldered to L4 which holds the diode MUST be mounted so that it is at right angles to the wall of the box to prevent the diode shearing when the bolt is tightened. This is not the same as making it parallel to the bottom of the box, as the walls of a diecast box always slope.

G4MBS comments that failing to observe this results in having a snap varactor from which there is no recovery!

Mounting the diode

The OBA brass screw has a 1/16" hole to take one end of the diode. The other end fits in a 1/16" hole in the 8mm dia, 5mm long brass rod which is soldered to L4.





You can align the axes of the two holes with a dummy diode, made from a spare OBA bolt whose end is turned down to 1/16"dia. This can then be screwed through the diecast box and into the hole in the other brass rod. When these are well seated, solder L4 to C6 and L3/L2.

Then remove the dummy bolt, and insert the bolt holding the diode. Gently tighten the bolt, and lock it in place with the locknut.

7.2 10GHz Band Pass Filters

7.2.1 Post Coupled Filters

Filters for 3cm

The growing interest in narrow-band operation on 3cm has led to a demand for filters to attenuate unwanted harmonics generated in multiplying from lower frequencies. G8APP has designed two filters to cover the top half and the bottom half of the band. As is shown by Fig 2, eight posts are fitted across the guide in precise positions, and two tuning screws are used to set the frequency. The filters have an 80MHz equal ripple bandwidth, a 3dB bandwidth of 120MHz and an insertion loss of 0·3dB. Frequencies 1GHz away are attenuated by about 40dB.

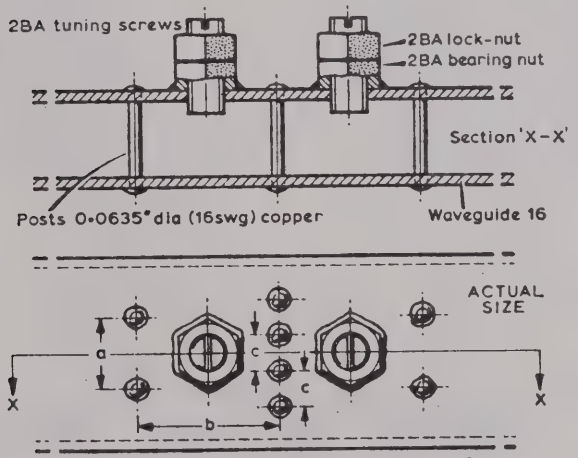
Although this type of filter is relatively tolerant, the posts should really be positioned within a few thousandths of an inch. Ideally, a milling machine would be used and the subsequent alignment done with elaborate test equipment. However, G3WJG has demonstrated that perfectly acceptable filters to this design can be made by marking out using an accurate rule and a magnifying glass, drilling with a diy drill and stand, and aligning by means of a klystron or Gunn oscillator, a variable attenuator and a mixer diode as a detector.

The method of assembly used by the writer is as follows:

1. Straighten the copper wire by stretching. Cut to length and squeeze one end to stop it slipping through the holes in the waveguide.

2. Fit the wires and the flanges, and hold the 2BA bearing nuts in position using chrome-plated screws. Using a flux which leaves a residue soluble in water (eg Bakers Fluid), and preferably a small gas flame, solder the flanges and also the post joints on the *underside* of the guide only, which will prevent solder running into the inside of the guide.

3. When the solder has solidified, invert the filter and place a piece of wet rag over the post joints already soldered. Solder the 2BA bearing nuts and the remaining post joints, again from the underside, allow to cool and then wash well in running water.



Frequency GHz	a	b	c
10·0 — 10·25	0·375"	0·730"	0·198"
10·25 — 10·50	0·370"	0·707"	0·197"

Fig 2. Waveguide filters for 10 to 10·5GHz

extracted from RadCom

7.2.2. Iris Coupled Filters

Designs for 10GHz bandpass filters

These filters for 10GHz have 3dB bandwidths of 20 or 60MHz, as is shown by the measured response curves given in Fig 1. G3JVL, who designed the filters, uses the former in the local oscillator chain of a receiver to reduce the noise generated at signal frequency which would otherwise degrade the receiver performance. Even if a relatively low i.f. is used, significant rejection of noise can be obtained; for example, at 30MHz the rejection is 27dB. The 60MHz bandwidth filter is used to eliminate effectively the second channel response of the receiver.

As is shown in Fig 2, the filter consists of a length of waveguide 16 in which iris plates are used to define three resonant cavities which are coupled by centrally placed holes of specified dimensions. The basic design frequency is 10·5GHz, but the tuning screws fitted may be used to tune the filters down to 9·5GHz at least. Matching screws are fitted at each end of the filter as necessary to tune out any mismatch with the external circuitry.

In order to maintain the Q of the filter, it should preferably be made entirely from copper. Brass should only be used if it can be copper- or silver-plated after assembly. The tuning screws can be of brass only if little of each projects into the cavities, otherwise threaded copper rods should be used. The matching screws can be of brass, or even cadmium-plated steel if little projects into the guide.

The filter can be constructed as follows:

(a) Scribe deeply into the top and bottom broad faces of the guide grooves corresponding to the position of the irises.

(b) Using a fine saw such as a junior hack-saw blade, first cut slots at each corner and then use these as a guide to extend the cuts across the full width of the top and bottom walls of the waveguide.

(c) Fit the iris plates in their correct positions, and fix them in place by bending the corners of the iris plates. Jig the tuning and matching screw bearing nuts (if fitted) using chromium-plated screws. Fit any flanges required and solder all joints in a single operation using a small gas flame. Alternatively, solder only the joints on the underside of the filter, invert it and solder the second set while the first is cooled by small pieces of tissue soaked in water. This technique prevents any solder entering the cavities, which would reduce their Q. It is safer to use plain solder with a flux that can be washed away (eg Baker's Fluid) rather than to risk organic residues being left behind. Note that it is not necessary for the iris plates to make good contact with the side walls.

It is most unlikely that the filter can be aligned simply by connecting an rf source at one end and a detector at the other. A simple technique is to poke the inner of a short length of coaxial cable down through one of the matching screw holes between the filter and the rf source, and to measure the rf output using a diode connected to the other end. Adjusting the tuning screws will produce peaks or dips at resonance, after which the filter can be aligned in transmission. A more formal technique is to connect one end of the filter to a matched load, and the other end to an rf source via a directional coupler. With the filter off-tune, the amount of reflected power will be high, but this will decrease as the filter is tuned. When the reflected power approaches zero, the filter can be finally adjusted in transmission.

If required, the frequency of the filter can be raised by fitting large (OBA or M6) screws centrally through the side walls of each cavity.

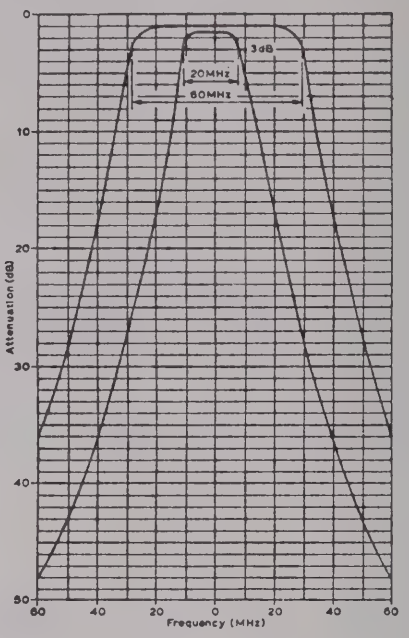


Fig 1. The response curves measured for the filters

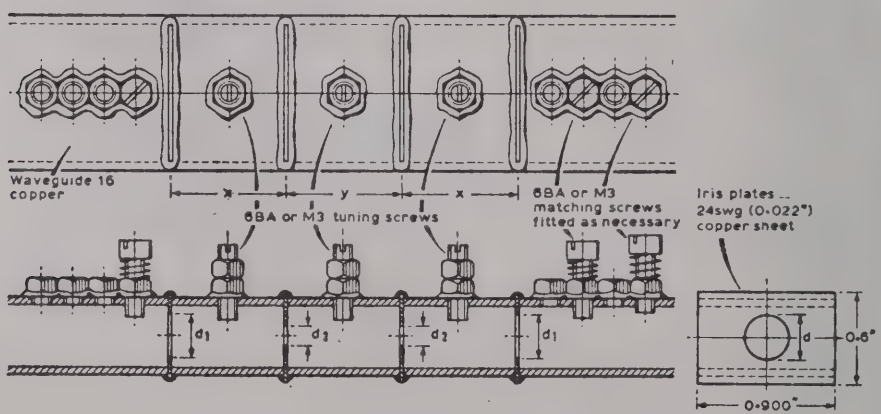


Fig 2. The construction of the filters. The dimensions are given in the table

Filter 3dB bandwidth MHz	Dimensions (mm) X	Y	d_1	d_2
60	17·8	18·2	9·0	4·3
20	18·2	18·7	6·2	2·95

extracted from RadCom

7.3.1 Original design (waveguide components)

The G3JVL image-recovery mixer/transmit converter for 10GHz

Readers will no doubt be aware of the regular tropo-scatter tests being carried out between Oxford and Hayling Island. G3JVL has supplied details of the mixer which is in use by both stations as a receive and transmit converter.

The unit consists of a mixer diode mounted between two filters, one on the local oscillator frequency (10,224MHz), the other on the signal frequency (10,368MHz). The local oscillator filter is designed for maximum rejection of local oscillator noise, and is therefore a very narrow band device, while the signal filter is designed primarily for low loss, and is therefore somewhat broader. The construction of these filters was described in detail in "Microwaves" (October 1977), so is not reproduced here in full detail. The relevant dimensions, ie the spacing and diameter of the iris holes are given here in Fig 1.

Both the multiplier and mixer diodes are mounted using Gunn diode type posts and discs, as described in *Radio Communication* (June 1978, p493). The matching into the multiplier diode can be the same as that described in "Microwaves" (March 1976). The diode requires, typically, 0·5–1W drive at 384MHz. G3JVL has an alternative matching network which allows the use of a lower drive level, and this will be published in due course.

The mixer diode is a GaAs Schottky barrier type, of AEI manufacture. Devices of the same type by other manufacturers are probably also satisfactory. The 144MHz i.f. output is fed by a suitable matching network to a low-noise preamplifier. The matching will depend on the exact diode current used, and is best set up using an automatic noise figure meter.

The tuning up procedure of the mixer unit needs to be done carefully if optimum results are to be achieved. Initially monitor the mixer diode current with 384MHz drive applied to the multiplier diode, and maximize this by using the local oscillator filter tuning screws and the tuning screws between the multiplier diode and the filter. Next, remove the 384MHz drive and apply rf at 10,368MHz to the input of the receiver. Tune the signal filter and all the rest of the matching screws for maximum current, reducing the rf input as necessary to keep the mixer current at the required 1·5mA. Finally, re-apply local oscillator drive and optimize all tuning and matching adjustments for best signal-to-noise ratio using either a weak-signal source or, preferably, an automatic noise figure meter.

When optimized, the noise figure should be around 7dB, allowing for a 1·5dB noise figure i.f. preamplifier. This is a very good performance indeed, and will only be bettered by the use of an rf preamplifier.

G3YGF has demonstrated the efficiency of this design as a transmit converter, by feeding a few milliwatts of 144MHz ssb into the mixer diode. About 1mW p.e.p. output at 10,368MHz can be achieved, which is adequate for local contacts or for driving a twt amplifier. □

Correction to G3JVL image recovery mixer

A small error occurred in the description of the G3JVL image recovery mixer described in *Microwaves* (January 1979). This was that 384MHz was quoted as the drive frequency to the step-recovery diode. This should have been 378·667MHz, to give a 144MHz i.f. A suitable starting crystal frequency for this would be 94·6667MHz; it is not recommended to start at a lower frequency, for local oscillator noise considerations. □

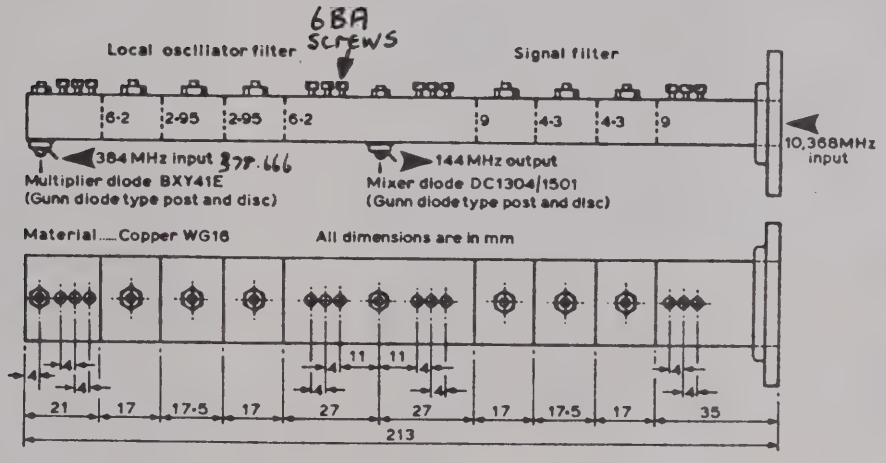


Fig 1. The G3JVL image-recovery mixer

extracted from RadCom

7.3.2 page 1

Further information on the G3JVL 10GHz transverter

by CHARLES SUCKLING, G3WDG*

Introduction

Since the design of the G3JVL 10GHz narrow-band transverter was published in *Microwaves* (January 1979) much interest has been shown in this device. A large number have now been built (including four by the author) and much experience has been gained in both their construction and alignment. What follows is an attempt to expand somewhat the original description in the light of this experience.

Construction

Two modifications to the design can simplify the tuning-up procedure considerably. These are the construction of the mixer and step-recovery diode multiplier as two separate units, and a minor modification to the arrangement of the matching screws. These modifications are shown in Fig 1. With this arrangement it has been found convenient to use the G8DEK multiplier, exactly as described in *Microwaves* (March 1976). The overall length of the multiplier, from the front of the flange to the rear short-circuit, was 29mm in the author's versions. An attractive feature of this design is the built-in attenuator pad in the input circuit, which results in very stable operation due to the high degree of suppression of interactions between the multiplier and the output stage of the 378·666 MHz exciter. The price for this, of course, is an increased drive requirement (1·5–2·5W) of the multiplier, but this is more than outweighed by the improved stability. A point not made in the original description of the multiplier is that the diode should be mounted with its heatsink end (ie that end remote from the flange) in the 2BA screw.

The original mixer design also called for the use of copper screws for cavity tuning to minimize losses. In practice it has been found that the use of brass screws instead of copper ones causes hardly any degradation in noise figure, and so it would seem not worth the extra trouble involved in manufacturing copper screws.

The construction of the mixer can be made as easy as possible if a careful procedure is followed. There are several "irreversible" stages, which must be done in the correct order to avoid later trouble. First saw off a length of waveguide and roughly square off its ends by filing. Mark out the positions of the iris slots carefully (aim for ±0·1mm accuracy). The slots may then be cut out using a junior hacksaw (with a new blade!), starting at each corner and working inwards towards the centre to maintain the accuracy of each cut. All holes are then marked out and drilled 2·3mm (6BA tapping). The centre hole must be drilled carefully, using a vertical drill, through both walls of the waveguide. The upper hole is then opened out to 3/32in and the bottom to 4mm (2BA tapping). This hole is then tapped 2BA, and the 18 holes for the tuning and matching screws tapped 6BA.

Next, the whole assembly is carefully deburred by running a file through the waveguide. This tends to push metal back into the holes and slots, which are then cleared using the hacksaw and appropriate drills and taps. Further filing and cleaning out should be repeated until all traces of burrs and loose metal are removed. The final deburring operation is to run a 4mm drill through the 2BA tapped hole just into the 3/32in hole, and rotate slowly by hand to cut a slight chamfer on the inside wall of the waveguide. This ensures a flat surface for the diode post to be mounted against, with no burrs remaining to puncture the insulation.

Details of the mixer diode mount are shown in Fig 2. After manufacturing the component pieces of this, deburr the surfaces which are to be mounted against the walls of the waveguide, by rubbing them on emery paper placed on a flat surface. The diode mount should then be assembled and checked for insulation, and for alignment using a scrap diode. For this operation the 2BA nut should be screwed up against the waveguide wall by hand. Should the diode not quite fit, the upper hole may be filed slightly so that the assembly may be moved to the correct position. A very useful tool for checking the alignment of diode mounts of this type can consist of a 2BA screw with a 1·8mm diameter pin turned on one end. This is used in place of the 2BA screw and scrap diode referred to above, and will positively assure that the mount is correctly aligned. The importance of this cannot be too highly stressed, as a mount which is out of alignment will result in broken diodes (an expensive experience!). When the diode mount has been checked, it should be dismantled. In order to obviate problems later on, ensure that the 0·3 by 0·150in piece will pass freely through the 2BA nut and the 2BA tapped hole in the waveguide. If not, open these out slightly with a suitable drill.

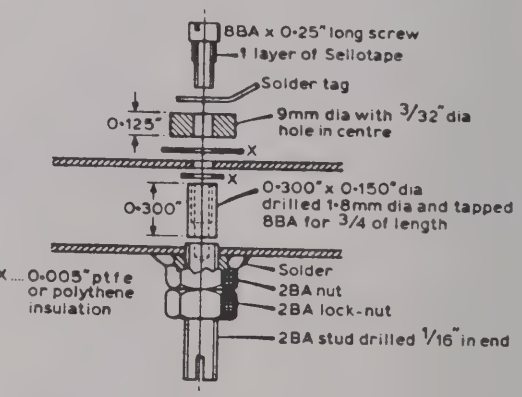


Fig 2. Arrangement of the mixer diode mount

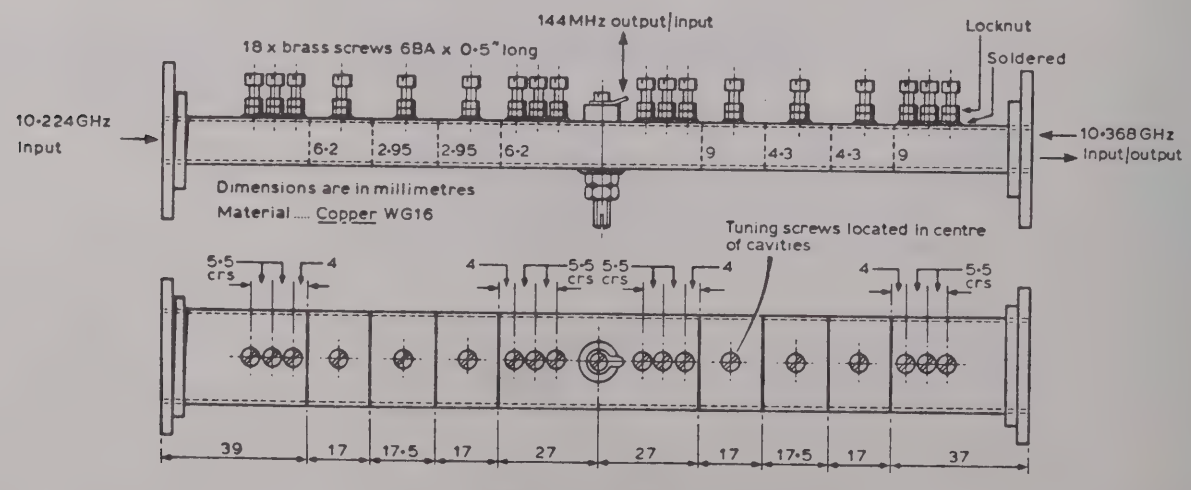


Fig 1. Constructional details of the modified G3JVL 10GHz transverter

extracted from RadCom

7.3.2 page 2

The next items to be made are the iris plates. These are manufactured form 0·9 by 0·5in pieces of 0·024in copper sheet which have been carefully filed to size; the 0·5in dimension, not being as critical as the 0·9in dimension, which should be in the range 0·895–0·900in. The iris holes are drilled centrally in the plates; the diameters of these are given in Fig 1. After deburring and flattening the plates between two smooth pieces of metal in a vice, they may be assembled into the waveguide. The 6BA nuts which are to be soldered to the waveguide are jigged in place using cadmium-plated or stainless steel screws. Old steel or brass screws which are sufficiently corroded not to take solder easily can also be used.

The assembly is then soldered, using a hotplate to provide the heat—direct heating with a flame is much less satisfactory. A suitable hotplate could consist of a piece of 1/8in aluminium plate on a domestic cooker ring. Perhaps the most critical part of the construction of the mixer is this soldering stage, as too much solder or flux inside the cavities can introduce excessive losses and/or detuning. Thus the cleaning of all parts to a bright finish before soldering is necessary. If all the parts are clean it should only require the application of a small amount of solder to each side of the irises to solder them. However, be careful that no gaps remain after soldering. The author has found 18swg Multicore solder to be ideal for this operation. Speed is essential, or the copper oxidises and becomes more difficult to solder. In this context the temperature of the hotplate should be only just high enough so that the solder runs freely.

When all the 6BA nuts and upper sides of the irises have been soldered, the waveguide is turned on its side and the bottom iris joints soldered, applying the solder to the tops of the joints. The 2BA nut is then jigged in position (using a suitable screw, as described above, and soldered. Next, the flanges are fitted in position, having been preheated on the hotplate, and soldered. A small amount of waveguide should be left sticking out of the flanges, to be removed later by rubbing on a sheet of emery paper laid on a flat surface, which then leaves a well-finished surface. Remove the assembly from the heat and, when it has cooled to just below solder melting point, solder several solder tags to the top and bottom walls of the waveguide, using an iron. These are to act as anchoring points for the i.f. preamplifier housing.

When the unit has fully cooled, remove all the jigging screws and run a 2BA plug tap through the nut and into the waveguide. The tap should be inserted vertically upwards and withdrawn vertically downwards to reduce the chance of metal dust falling into the cavity. This tapping operation ensures that the 2BA diode-mounting screw runs freely in the thread. Next, the diode mount may be reassembled. This *must* be done in the following way to avoid losing pieces inside the waveguide, which are then very difficult to retrieve (as the author knows to his cost—twice!).

Pass a 1in-long 8BA screw through the top hole and out through the 2BA nut. Place the small insulating washer over the screw, and screw on the 0·3 by 0·15in post hand-tight. Then pull this piece through the 2BA nut up against the inside wall of the waveguide, and insert a 2BA screw or, preferably, the alignment tool mentioned above, and screw in until the post is pushed firmly against the wall. The 8BA screw is then removed and the top part of the diode mount assembled. The 2BA screw is removed and replaced by the 2BA diode-mounting screw complete with diode. In order to avoid dropping the diode inside the cavity, the screw and diode are held with the diode uppermost and then fitted vertically upwards into the waveguide. The screw is tightened by hand so that the diode is held in place, and the locknut tightened up. Diode removal is achieved by slackening off the locknut, unscrewing the 2BA screw one turn, tapping the screw gently sideways to release the diode from the upper post, and.then unscrewing downwards. The mount itself can be dismantled, if required, by reversing the assembly sequence.

Initial tuning-up

As with the construction, the alignment of the unit is much simplified by the adoption of a logical procedure. The first step is to connect a multimeter to the diode, via short leads to reduce the risk of transient pickup, which could damage the diode. Next, the waveguide test bench shown in Fig 3 should be assembled from the station 10GHz (wideband) equipment. The directional coupler is orientated so that the detector current is a monitor of reflected power, ie the current should be very small or zero with a matched load connected in place of the mixer, and several hundred microamps with the mixer in place (and off tune). Set the Gunn oscillator to 10,368MHz and connect the signal input end of the mixer to the directional coupler. The procedure is to then adjust the first filter tuning screw (ie that tuning screw nearest to the directional coupler) until a sharp change is seen in the level of reflected power, indicating resonance. The three matching screws in front of the filter should not be inserted at this stage. Leave the tuning screw at the position of resonance, and adjust the second cavity tuning screw for a similar effect. Adjust the third tuning screw likewise, and then repetitively adjust the three screws until no reflected power is indicated. Recheck constantly during the alignment that the Gunn oscillator is still on frequency, as it is possible that its frequency may be affected by the tuning of the cavities.

At this point there should be an indication of a few microamps on the meter connected to the mixer diode (referred to below as the "mixer current"), indicating some transmission of power through the filter. If no current is indicated, slowly insert a matching screw into the hole nearest to the signal filter, on the mixer diode side of the cavity. As the reflected power is seen to rise, readjust the tuning screw in the adjacent filter cavity for minimum reflected power. As this procedure is continued with increasing penetration of the matching screw, at some point some mixer current should be seen. When some mixer current has been obtained, readjust the filter tuning screws, and insert matching screws on either side of the signal filter and experiment with their penetrations until maximum mixer current is obtained. As all the screws tend to interact with each other to some extent, it is necessary to go several times round the "loop" until the optimum settings are found. A mixer current of several milliamps should be achieved using a 5–10mW Gunn oscillator.

Lock the screws in position, readjusting if necessary to maintain the tuning, remove the mixer from the directional coupler and reconnect the other way round. Reset the Gunn oscillator to 10,224MHz and repeat the above procedure for the local oscillator section of the mixer. Do not interfere with the tuning and matching screws associated with the signal filter. Since the local oscillator filter is somewhat narrower in bandwidth than the signal filter, it is rather more difficult to adjust, and greater care will be necessary.

The step recovery multiplier should be aligned next. Connect its output to a detector and, on application of drive at 378·666MHz to the multiplier, some current should be seen in the detector. Maximize this by repeated adjustment of the tuning screw, the two trimmer capacitors and the variable resistor on the multiplier. When several milliamps of detector current have been obtained, the multiplier may be transferred to the mixer unit, whereupon some mixer current should be observed. This is maximized by repeated adjustments of the multiplier and the local oscillator filter and its associated matching screws. The mixer current should then be at least 1mA, and possibly as high as 5mA. The final tuning-up stages require the use of the i.f. preamplifier, and will be discussed later. Before proceeding further it is best to remove the mixer diode to avoid any chance of damage by transients during subsequent soldering operations.

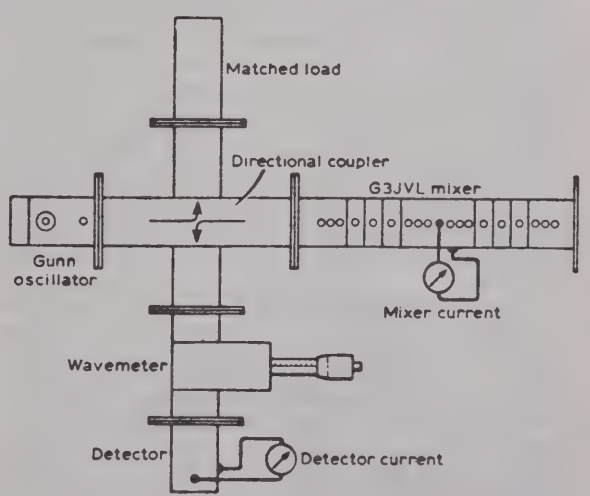


Fig 3. Waveguide test set-up for aligning the mixer

7.3.2 page 3

A suitable i.f. preamplifier

No information was given in the original *Microwaves* item concerning i.f. preamplifiers. Since the noise figure of a mixer is partly dependent on the noise figure of the subsequent receiver, a low-noise i.f. preamplifier is vital. For optimum performance this should be mounted on the waveguide assembly.

The circuit and suggested layout of a suitable preamplifier are shown in Figs 4 and 5. Note that if receive-only operation is required the relay may be omitted.

Final alignment

The final alignment of the mixer requires the use of an automatic receiver alignment aid, such as that referred to in *Microwaves* (October 1979). First adjust the 20pF trimmer in the preamplifier for maximum receiver noise. Then connect up the alignment aid and reset the trimmer for best noise figure. Adjustments may then be made to all the tuning and matching screws for best noise figure, although these are somewhat interactive, and it may take time to reach the optimum settings. Once an optimum has been found the output power on transmit can be checked by applying sufficient drive (5–20mW) at 144MHz to approximately double the mixer current. This drive may be obtained from the station 144MHz transceiver via a suitable attenuator (*see Microwaves* March 1979) with an adequate power rating. The settings for maximum transmit power and best noise figure do not coincide exactly, and it will be necessary to compromise if real transceive operation is intended. However, if it is expected to use the unit mainly as a receiver, then optimize for best noise figure only.

Additional information

The original *Microwaves* item on the G3JVL mixer specified a BXY41E diode for the multiplier and an AEI DC1304 GaAs Schottky diode for the mixer. The former device is no longer readily available, while the latter is somewhat expensive. Some cheaper, more readily available diodes manufactured by Microwave Associates have recently been tested, with very satisfactory results indeed. A suitable replacement for the BXY41E is the MA44150, while the MA40150 and MA4E024 devices have performed very well as mixers, both with noise figures of 7·5dB. This is not quite as good a performance as that measured by the author using the DC1304C device, which had a 6·5dB noise figure, but is still very acceptable. These noise figures refer to the overall system using the i.f. preamplifier described above feeding a 3dB noise figure 144MHz receiver.

If full transceive operation is intended it will be necessary to add a coaxial relay, switched by the 144MHz transceiver ptt line to switch the transceiver from the receiver output socket to the transmitter input socket, as shown in Fig 6. The value of attenuation required in the transmit line should be such as not to overdrive the mixer. Suitable starting values are 20dB for a 3W output 144MHz exciter and 25dB for a 10W exciter. Extra attenuation should be inserted until the 10GHz output just starts to decrease; the attenuator can then be rebuilt to that value. It is important to ensure that the resistors comprising the attenuator will handle the power safely, since any failures could result in a large drop in attenuation with the consequent risk of damage to the mixer diode.

Little has been said concerning the 378·666MHz exciter. This is of course a crucial part of the system, and careful attention must be paid to its design in several respects. The most important of these concerns the noise generated by the crystal oscillator. A noisy oscillator can seriously degrade the overall noise figure; this problem beset early experimenters on 10GHz narrow-band. Also, the oscillator must be extremely stable, implying the use of high quality crystals as well as a good circuit. An exciter designed specifically for use with this mixer is currently being developed, and details of this will be published in due course.

In use, few problems have been found with this design. One thing which it specifically dislikes is water finding its way into the cavities, so operation in wet weather requires careful attention to be paid to waterproofing the unit. Most constructors mount the mixer, together with the changeover system and 378·666MHz exciter, in a large diecast box. Also, some detuning of the cavities can occur over extremes of temperature, usually resulting in low mixer current. The first screw to adjust in this event is the local oscillator filter tuning screw nearest to the mixer. Do not attempt to move two screws at once, or the mixer may well go irrecoverably out of alignment. Some more cautious operators have been known to take automatic noise figure optimization equipment out with them in the field, but this is not usually necessary.

Finally it should be recognized that this design is most definitely not for the beginner on 10GHz, both in respect of its relative complexity and the absolute necessity for test equipment, albeit homemade, for tuning it up. □

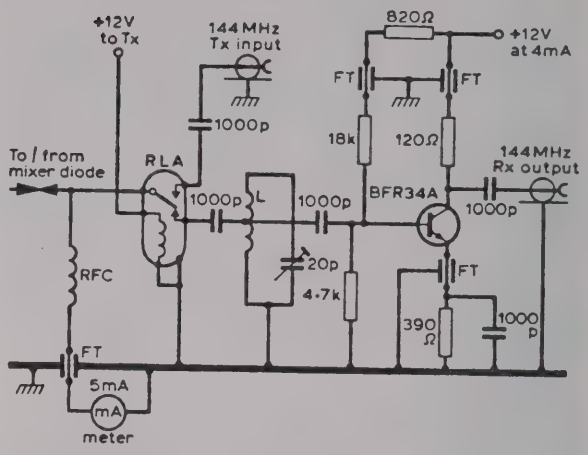
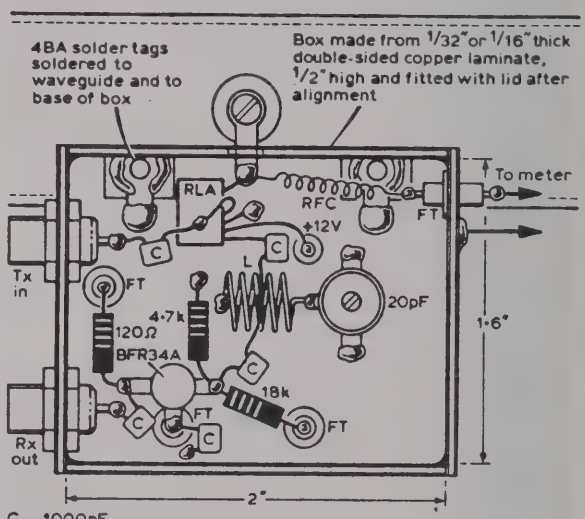


Fig 4. Circuit diagram of a suitable i.f. preamplifier and changeover system. All 1,000pF capacitors are miniature ceramic plate types. L: 5t 18swg, 0·55in long, 0·25in id, tapped at 2·5t. RFC: 20t, 26swg, closewound 0·15in id. RLA: RS type 349 591 or equivalent. FT: 1,000pF feedthrough capacitors.



C.... 1000pF
The 390 and 820Ω resistors are mounted on underside of the box

Fig 5. Layout of the i.f. preamplifier

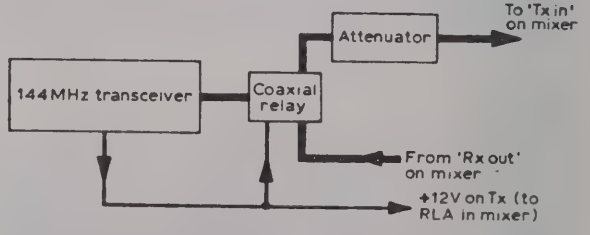


Fig 6. Block diagram of a complete 10GHz transceiver

Tapered Section Mixer - G3JVL/G3YGF

The original version was published in RADCOM (see April 1980) and since then G3JVL has produced a design for a modified mixer cavity, which is reproduced here. In this version

The original version was rather critical to align, and had many interactive matching screws. Details of the new version are given in Fig 1 a and b. In this version the mixer diode is mounted in reduced height waveguide that matches the diode's impedance. A tapered transition between the reduced height guide and the filters makes the matching very much less critical.

The coupling from each filter to the mixer cavity is made adjustable by using a new screw just inside the last filter cavity which effectively alters the size of the coupling hole in the iris plate. Since this screw is inside the filter cavity, it only affects the matching at the filter frequency, and so it does not affect the matching at the other filter frequency . It has effectively removed most of the interaction problems encountered during alignment.

This design has given results as good as the original, but without needing the six matching screws in the mixer cavity. It may be worth leaving provision for some matching screws there, as it may enable a dB or so improvement to be won. Several different ways of making the tapered section are suggested in Fig 1b.

This design is very highly recommended, as it is far easier to align, and its performance is more consistant and reliable.

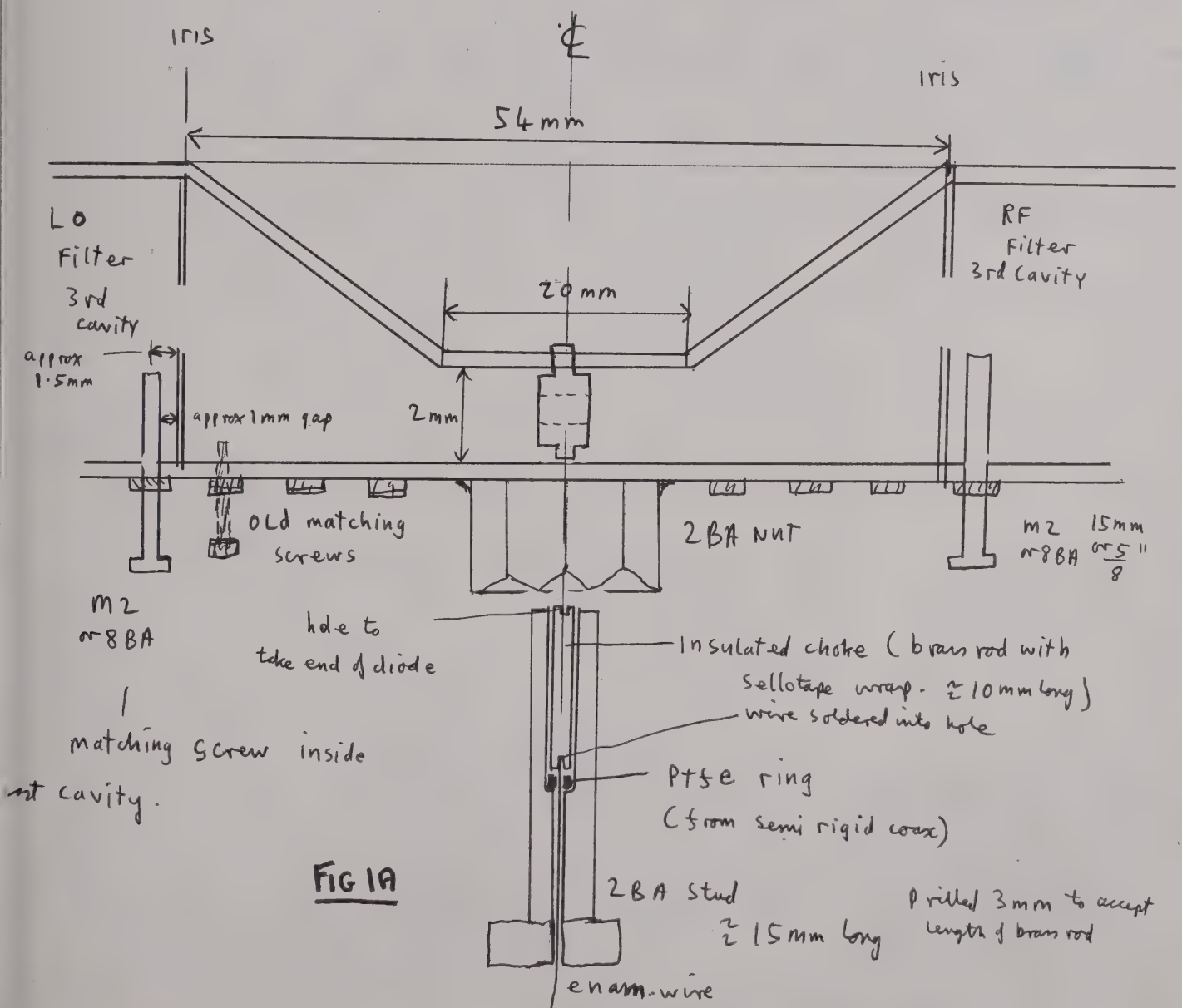
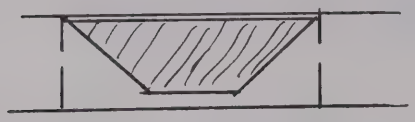


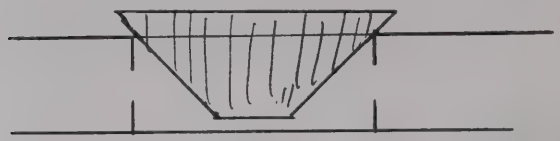
FIG 1A

7.3.3 page 2

1) Tapered block inserted in waveguide before soldering



2) Tapered block inserted through top of gui
(top face of guide removed.)



3) (Avoids machined block)

FIG 1B

shaped notch cut in waveguide (Through both side walls, and the top face removed. Copper foil is then laid along the trough, to form the top wall of the tapered waveguide section.

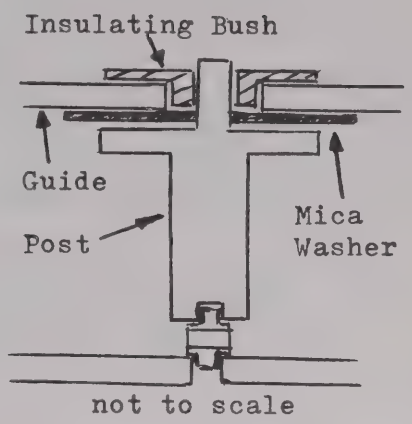
7.3.4 Improvements to multiplier and mixer diode mounts

Leonard, GM4HFM, has come up with some suggestions to improve the mounts for multiplier and mixer diodes in narrowband rigs.

SRD and Gunn diode mounting posts:

There are often problems when using thin PTFE or Sellotape as the insulating washer, as it can puncture and short out the diode. An alternative insulator is a mica washer that is normally used for mounting power transistors - these already have holes drilled in them, and can be adjusted in thickness by splitting the layers of mica apart with a needle.

It is also possible to use the plastic insulating bushes to keep the post central and insulated where it passes through the waveguide wall.



The tapered mixer assembly :

As an alternative to forming the tapered section with copper foil soldered in the guide, Leonard has suggested bolting a tapered block of brass or copper in place as shown below. This also makes it possible to use a socket for the mixer diode connection. The block is inserted through a hole in the upper face of the guide and held in good contact with the walls using 6BA bolts in the sidewalls. It is then very easy to drill and tap the block to accept the matching screws.

The advantage of using a socket mounted on the guide, is that it can be used to support the parts of the choke and diode connection. In some of the other designs either the assembling is rather difficult, or the insulation is likely to puncture after a while. Either of BNC or SMA sockets could be used. The choke is made from a section of low impedance transmission line formed by the large diameter inner rod, insulated by a layer of Sellotape.

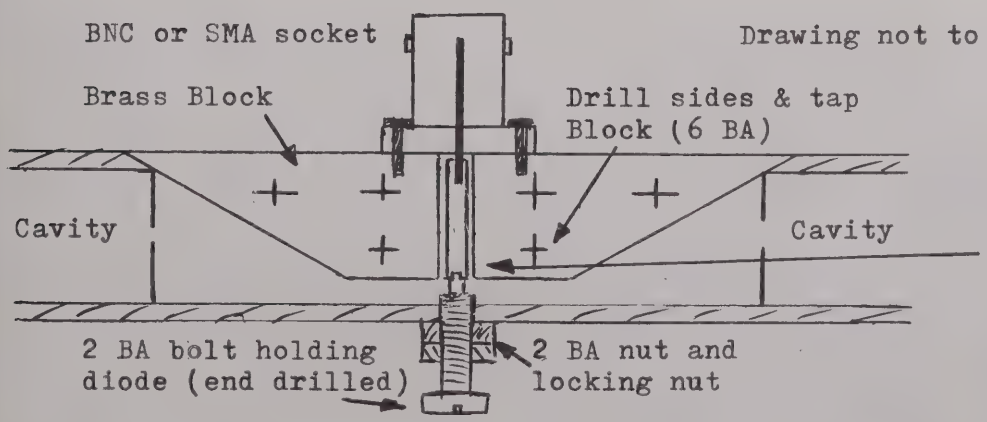
The diameter of the rod is not critical, somewhere between 5 - 8 mm, but the gap between the rod and the hole in the block should only be sufficient to allow the rod to be wrapped in one or two layers of Sellotape.

The characteristic impedance of this rod can be calculated from the following expression:

$$Z_o = \frac{\sqrt{\frac{\mu_o}{\epsilon_o \epsilon_r}}}{2\pi} \log_e \frac{b}{a} \quad \text{ohms}$$

where μ_o = permeability of free space, 4×10^{-7} H/m
ϵ_o = permittivity of free space 8.85×10^{-12} F/m
a = diameter of rod
b = diameter of hole
ϵ = dielectric constant of the insulator

For $\frac{1}{4}$" rod with $\frac{1}{4}$mm gap around it filled with $\epsilon = 2$ then Z_o works out at 3.2 ohms.


BNC or SMA socket
Drawing not to scale
Brass Block
Drill sides & tap Block (6 BA)
Cavity
Cavity
2 BA bolt holding diode (end drilled)
2 BA nut and locking nut
The slug is drilled at bottom for diode and at top to solder to inner pin of skt. Hole in brass block sufficient for rod and insulation.

Alignment of G3JVL Mixers - Several people have reported problems in getting these going, and it does seem to be one of those exercises where you need plenty of patience and faith, and a methodical approach. Most points are covered in G3WDG's article (RadCom, April 1980), but the following notes may prove of assistance:

1. The 'directional coupler' method of aligning the filters (by looking for a dip in reflected power as each cavity is tuned through resonance) is highly recommended. If an isolator or attenuator is available, this can be used immediately in front of the Gunn oscillator to prevent its frequency being pulled by the filter cavities.

2. After this, when aligning the filters in transmission, and looking for diode current in the mixer diode, an AVO can be used to monitor the reverse resistance of the diode. In this way, meter deflection can be obtained with power levels which would be too small to give a reading on a current range.

3. The G8DEK step-recovery multiplier (Microwaves, March 1976) is thoroughly recommended. It is important that the attenuator pad is mounted on the multiplier itself, rather than at the 376.6 MHz exciter end of a piece of cable. It's also more convenient if all the adjustables (i.e. 376.6 MHz matching capacitors, bias resistor, tuning screw and matching screws (if these aren't on the mixer)) are all on the top of the multiplier.

4. Once the 10224MHz LO has been fully optimised (i.e. step-recovery multiplier and LO filter), the multiplier and filter screws should need no further adjustment, and should not be touched (except perhaps the filter cavity screw nearest the mixer).

5. The 10368MHz signal-frequency filter is much broader than the LO filter, and is much easier to align.

6. The adjustment of the mixer matching screws is a compromise between max LO drive and best signal-frequency matching. G3WDG notes that on his rig, at high mixer current (~5mA), both a low receiver noise figure and over a milliwatt of SSB output are obtained - it therefore seems to be worth seeking as much mixer current from the LO as possible.

7. G3YJH suggests a method for aligning rigs where the step-recovery multiplier and mixer have been built from a single piece of waveguide: "By inserting an inch or two of insulating wire, which acts as a small antenna, through the tuning screw holes in turn, the signal can be tuned in on the station wideband rx tuned to 10224MHz. When the tuning screw of the preceding cavity is tuned through resonance, a 'blip' can be heard on the receiver as the signal peaks. By working through the cavities in this fashion several times, it is possible to peak the signal quite easily until sufficient mixer current is observed. The meter can then be used for the final alignment."

F6DLA suggests that the local oscillator in narrow band equipment should be slightly low of its nominal frequency so that the receiver covers from about 10367.900 to 10368.300 or so (using a usual 144 MHz tranceiver for the IF), so that signals on 10368.000 from CW transmitters using 96.000 MHz crystals can be found even in the worst case when both equipments drift the wrong way.

William warns that it is possible to listen to a spurious signal, rather than the main one, when listening to a 96 MHz derived 10368.000 transmitter on a receiver with a 144 MHz IF. The receiver normally listens on 10368.000 MHz, and with a 144 MHz IF has an image response on 10080.000 MHz. The 96 MHz crystal comes out on both these frequencies (96x105=10080, 96x108=10368). The same is true for a transverter on transmit - some power will also be produced on 10080 MHz as well as 10368 MHz. So particularly when testing over short paths, it is possible to hear the signal on the image frequency, despite the large attenuation that the 10368 MHz filter provides. Be careful not to align the equipment on the image by mistake!

7.3.5 page 2

G3PFR has some comments on the method of alignins the G3JVL mixer cavities by observins the reflected power. The first cavity should dip the reflected power, the second peaks it, and the third dips it again. This may be rather unexpected, but is in fact correct, and has been checked by usins a small probe into the next cavity throusht the hole for the tunins screw. He also emphasises that resonance will not be obtained unless the screws are very clean and a tisht fit in the nuts. This can be ensured by keepins a sentle torque on the locknut while the screw is beins adjusted. He wonders if the use of dielectric tunins screws misht overcome this problem, and is lookins for a source of ptfe 6BA screws. (I have just tried some nylon screws to tune the cavities and they do work, needins to penetrate slishtly further into the cavity, than brass ones, but they do increase the loss from about .8dB to 3.2dB for a 60MHz wide filter. PTFE may well be less lossy- G3YGF).

7.4 Comments on the VHF Comms Narrowband Tranceiver

VHF Comms Narrowband Tranceiver

G3UFW and G6GCI (Romsey) have been building this design and have experienced some problems in aligning it.
On receive it uses a varactor driven at 2268MHz and an IF of 1296MHz. There is very little filtering on the varactor output which means that several of the harmonics of 2268MHz reach the mixer. The receiver has two quite strong responses, one at 10368MHz ie

10368 - 4x2268 = 1296MHz

but also one at 10044 MHz ie

5x2268 - 10044 = 1296MHz

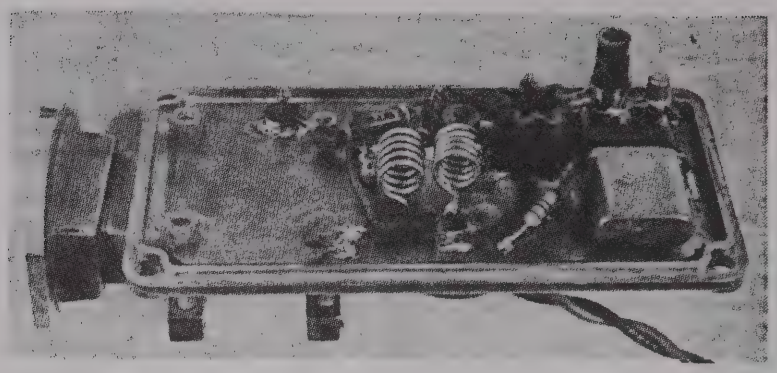
The tuning around the varactor should be optimised at 4x2268 ie at 9072MHz but the presence of several LO signals will increase the rx noise level somewhat, in addition to the unwanted response.
On transmit the varactor is fed with both 2268 and 1296 MHz and functions as a multiplier and mixer. This now means that the tuning around the varactor should be optimised at 10368MHz, and so the tuning is something of a compromise. With both inputs to the varactor there a number of outputs possible:

4x2268 + 1296	= 10368 wanted 5th order
8x1296	= 10368 unwanted 8th order
5x2268 - 1296	= 10044 unwanted 6th order
3x2268 + 3x1296	= 10692 unwanted 6th order

Some of these outputs are only 1 order higher than the wanted one and quite close to it in frequency, and so may be difficult to remove. It may be possible to inadvertantly align both the tx and rx up on 10044MHz if insufficient care is taken in setting up. In fact G3UFW's equipment had a significant amount of power output at 10044MHz. The 8th harmonic of 1296 cannot be filtered out but can be recognised by its very high tuning rate (8x normal).
Despite all this G3UFW and G6GCI hope to be qrv with narrowband soon. They would be interested to hear from others who may have had experience with this design.
NB The design is from 3/1980 and 1/1981 issues.

7.5.1 page 1

A crystal-controlled frequency marker for 10GHz



by C. W. SUCKLING, BSc, G3WDG*

Introduction

A characteristic of operating wide-band equipment on 10GHz is that by the time one multiplies the uncertainties in frequencies of both transmitter and receiver and in the pointing of the highly directional aerials normally employed, the number of permutations that need to be covered before contact is established can be very considerable. The more skilled operators tend to be more successful mainly because they can set their frequencies more precisely and can direct their aerials more accurately, and this can reduce very considerably the number of permutations that need to be covered.

The simple unit to be described generates a large number of signals at precisely known frequencies which can be used to calibrate wide-band receivers and transmitters accurately, thus effectively eliminating uncertainties from this source. Although the output power of individual harmonics is very low, signals are detectable with efficient receivers even with up to 45dB attenuation between the unit and the receiver. This means that the unit can provide a rough check on the sensitivity of a receiver; signals should be detected by even the most inefficient of receivers. The output power is more than adequate for use as a frequency reference for an afc system, but is insufficient for the unit to be employed as a signal source for tuning up aerials. The range of the unit as a transmitter is only about 10m when aerials of 15dB are used.

Constructional details

The circuit diagram and layout of the marker unit are shown in Figs 1 and 2. TR1 is a crystal oscillator on 48MHz, and this is followed by a doubler to 96MHz and by a mixer diode used as the final multiplier. TR2 should be fitted with a small heat sink and its emitter lead kept as short as possible, typically about ⅛in. Its output is via a bandpass filter to reduce 48MHz feedthrough. This vhf circuitry is mounted on a piece of single-sided pc board which is bolted to the inside of the lid of a 4¾ by 2¾ by 1in die-cast box. The final multiplier consists of a length of WG16, at least 2¾in long, which is closed at one end, with a 1N23 mixer diode mounted centrally 0·29in (7·5mm) from the closed end. The waveguide is clamped to the outside of the lid, connection to the diode being made using the inner of a Belling-Lee socket which passes through holes drilled in the pc board and the lid, as shown in Fig 3.

In setting up the unit, L1 is adjusted to produce the maximum voltage across R; C1 and C3 are adjusted and readjusted to maximize the diode current measured at TP1. The value of R is then changed as necessary to set this current to 20-25mA. Finally, the frequency of the crystal oscillator can be checked on a counter via the optional output connector or using, for example, a 144MHz receiver. Some useful performance parameters are given in Table 1.

Table 1. Performance figures for the 10GHz frequency marker

V supply	12V
I supply	25–40mA
$\frac{\Delta f}{\Delta V}$ (at 10GHz)	8kHz/V (change in final frequency as supply voltage is varied)
Rf level of 96MHz harmonics in range 10·0–10·2GHz	—75dBm (approx.)
Rf level of intermediate 48MHz harmonics in range 10·0–10·2GHz	—95dBm (approx.)

The choice of 48MHz and 96MHz as stage frequencies represents a compromise between generating a reasonable number of signals within the tuning range of most receivers, while minimizing the risk of confusion due to difficulty in identifying each of the harmonics. Other crystals may of course be used, but there is an obvious advantage in using "round number" frequencies, and especially those at such frequencies as 36, 54 and 72MHz which also produce harmonics at other amateur frequencies and are therefore more easily measured.

Using the frequency marker

The unit as described produces relatively strong signals at 96MHz spacings, with signals about 20dB weaker spaced every 48MHz. Note that no provision is made for modulating the output. This is unnecessary if the local oscillator of the receiver can be frequency modulated with a tone, which greatly assists finding signals. In calibrating a receiver it is essential that its local oscillator is already calibrated to within about 20MHz using a wavemeter. The self-calibrating wavemeter described in "Microwaves" (*Radio Communication* February 1973, p 106) is more than adequate for this purpose. The marker is preferably connected directly to the receiver via a variable attenuator, but alternatively it can be spaced from the receiver input by a few feet. With 20-30dB

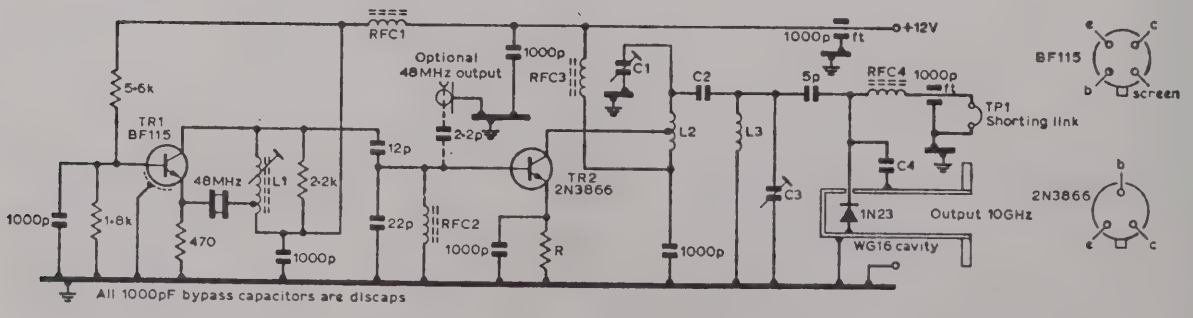


Fig 1. Circuit of the 10GHz frequency marker

extracted from RadCom

7.5.1 page 2

Components list

L1	10t 28g enam copper on 0·3in former tapped 1t from cold end.
L2, L3	5t 18g ¼in id ¼in long, L2 centre tapped.
RFC1, 3	2½t 28g enam copper on two FX1115 ferrite beads.
RFC2	2½t 28g enam copper on one FX1115 ferrite bead.
RFC4	10t 28g enam copper on one FX1115 ferrite bead.
R	68–100Ω, adjust during alignment to set 1N23 current to 20–25mA.
C1, C3	30pF trimmer: can be 20pF trimmer + 10pF fixed capacitor.
C2	Two lengths thin single-strand insulated wire twisted together for ¼in.
C4	formed by a 0·005in ptfe or polythene sheet between the end of the 1N23 diode and the waveguide wall, as shown in Fig 3.

attenuation, the receiver should detect weak signals which are harmonics of 96MHz only. If the attenuation is reduced to around 10dB, the 48MHz intermediate harmonics should be heard as weak signals, together with the now strong signals corresponding to 96MHz harmonics.

The number of signals heard will be more than perhaps expected because the receiver will almost certainly have no protection against second channel signals. Thus a receiver which is fitted with a 30MHz i.f. will respond to the relatively strong marker signals at 9,984MHz (104 × 96MHz) when its local oscillator is tuned to either 9,954 or 10,014MHz, and to signals at 10,080MHz (105 × 96MHz) when tuned to either 10,050 or 10,110MHz. The receiver will also respond to the relatively weak marker signals at 10,032MHz (209 × 48MHz) when the local oscillator is tuned to either 10,002 or 10,062MHz, and so on. The receiver local oscillator may therefore be calibrated precisely at several points, from which the (two) corresponding signal channels may be determined (provided that the i.f. is known accurately). If a larger number of calibration points is required, then either the i.f. may temporarily be changed, or a different crystal may be used in the marker.

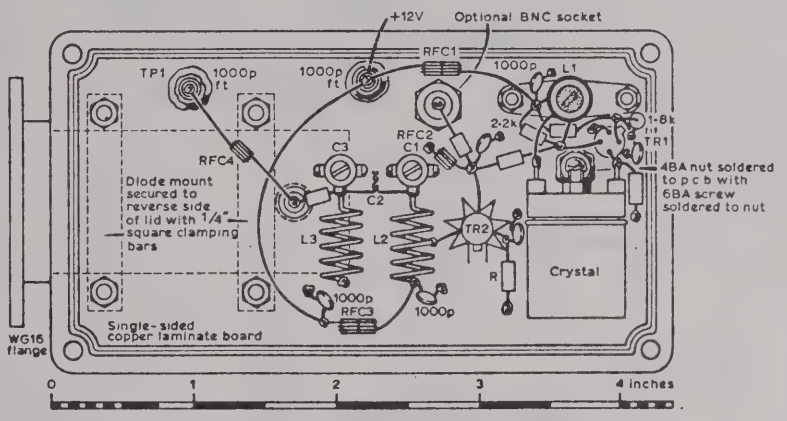


Fig 2. Practical layout of circuitry

The unit is small enough to be used "in the field" on a regular basis by simply waving it in front of the aerial. Alternatively, it could easily be built into the receiver, coupled by a 10dB directional coupler, to provide an instant check on its calibration, and in turn that of the transmitter.

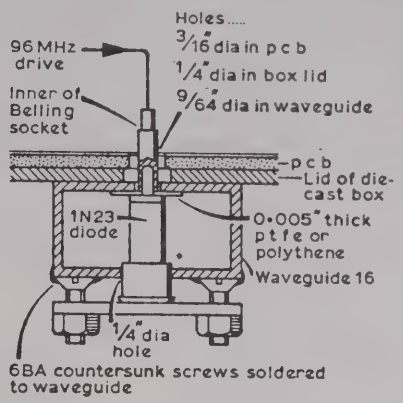


Fig 3. Section showing construction of the diode multiplier

Acknowledgement

The author wishes to acknowledge the help of G3RPE in the production of this article. □

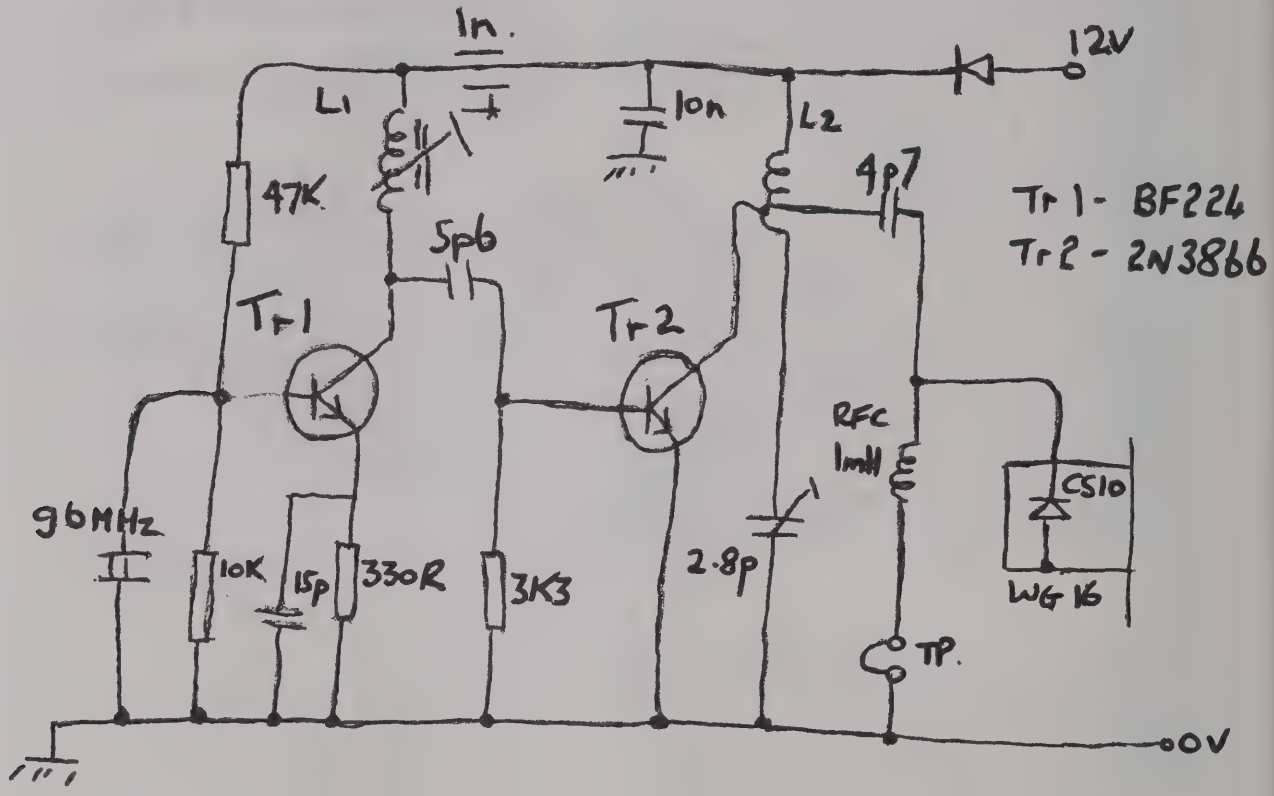
7.5.2 Marker Oscillator for 10GHz

G3ZIV has built the following circuit to produce weak markers on 10GHz. There is a similar design by G3WDG in Radcom but this one uses a 96 MHz xtal to drive the diode directly at 96MHz; the WDG version multiplied a 48MHz xtal to 96MHz before driving the diode. This one will probably give rather more drive to the xtal.

The diode used to multiply from 96MHz to 10GHz is any mixer diode in a normal waveguide mount. The drive to the diode should be such that about 20mA of diode current are seen in a milliammeter inserted in the link. Do not use a long coax lead between the output stage and the diode, otherwise the high capacitance may reduce the drive level. Ideally the circuitry and diode mount should be in the same box.

The harmonics on 10GHz are not very strong, but should be audible on a wideband rig when the source is within a foot of the receiver waveguide. On a narrowband receiver, it may be heard over a distance of several feet with open waveguide at each end.

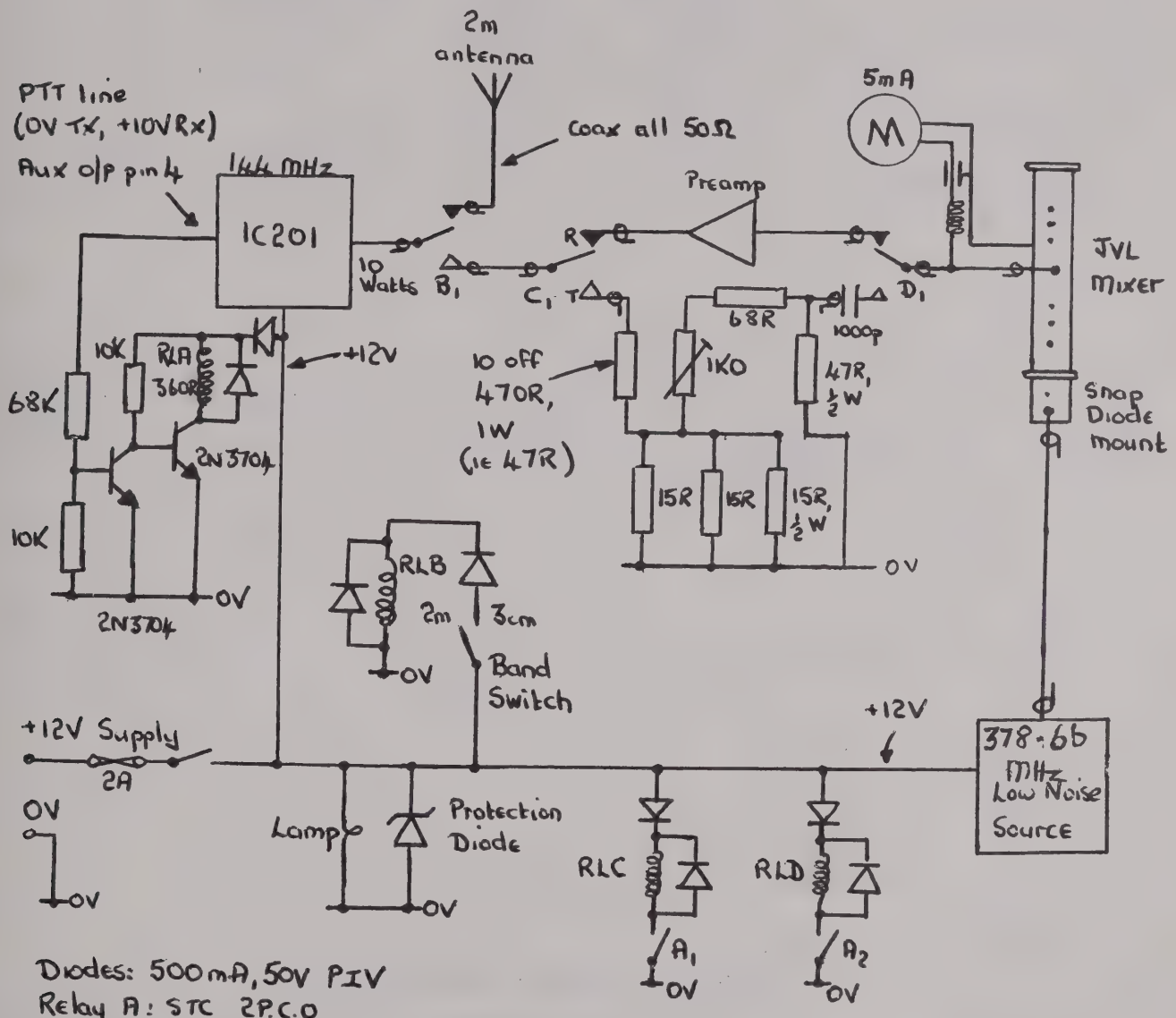
It is useful for checking the frequency calibration of wideband rigs, and the sensitivities of both wide and narrowband receivers. It is unlikely to be stable enough to calibrate frequencies on a narrowband receiver except over fairly short periods.



L1 - 7 turns of 18swg, 6mm dia. L2 - 4 turns of 16swg, 6mm dia.

Change-over System - G3AYJ

The change-over system below is as used by Dennis Morris, G3AYJ with his narrowband JVL mixer for 3cm operation. Besides the change over from receive to transmit on the mixer, the facility is provided to switch the 144 MHz tranceiver directly into an antenna for talkback. While it may still be necessary to retune when doing this, the operation of unplugging and plugging in leads is eliminated.



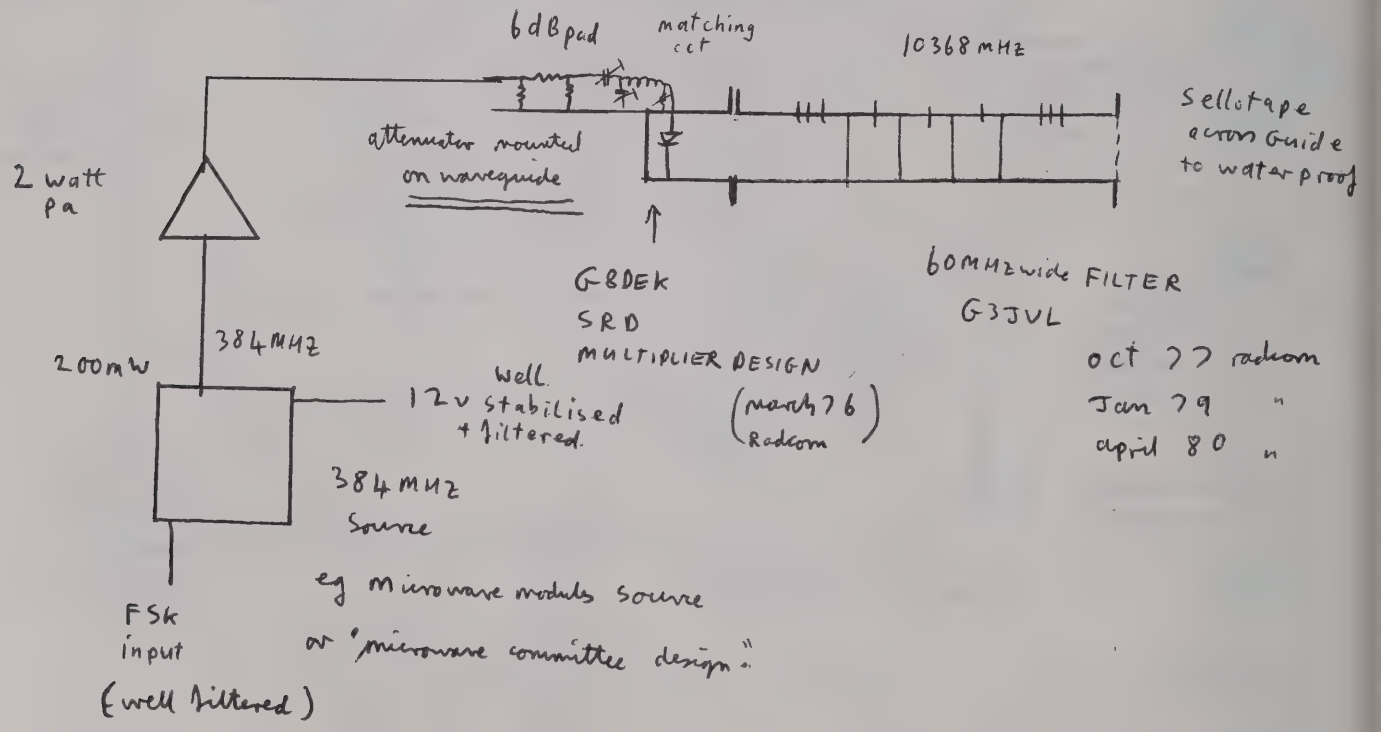
On transmit a power attenuator is switched into position when operating on 3cm, and a preamp is used on receive on 3cm. Note that the power attenuator will also be switched into position when operating directly on 2m, and also in the event of the PTT line being accidentally disconnected. It is not possible to have a 2m power output available without a supply voltage to the relay control circuits. The use of diodes in the positions shown is important for suppression of transient voltages.

Beacon Building: Not everyone appreciates just how useful beacons are - for checking receivers, for use as calibration references for pointing antennas, and (perhaps most important) long - term monitoring as part of propogation experiments. For these reasons, the microwave committee is very keen to promote beacon building, and would be prepared to help with design and hardware if at all possible. 10 GHz narrowband beacons (for which the design is fairly straightforward) would be particularly welcome. If anyone is interested, please contact G3YGF, G3WDG or G4CNV.
NB: It is fairly straight forward to combine wideband and narrowband sources into the same aerial, or to incorporate narrowband into an existing beacon, provided the antenna will work at 10368 MHz.

Do not be deterred by the prospect of having to build a high power multiplier chain for beacons. It really is not essential - 5 mW output is very easily obtained, and this is still potentially better than a wideband system running 100 mW as shown below:

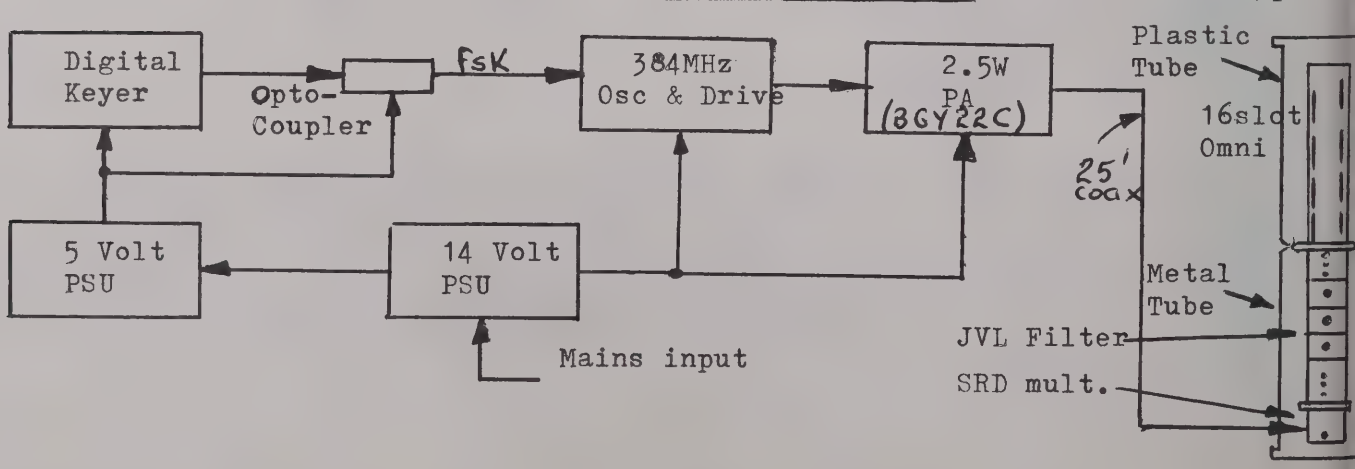
	Bandwidth	Power	Rx Threshold	
WB	300 kHz	100 mW	10 (FM)	
NB	3 kHz	5 mW	0 (CW)	
NB Advantage	+20 dB	-13 dB	+10 dB	= +17 dB

The design suggested below is useful either for a beacon, or a personal marker/CW source:



Details of G3AYJ's 10GHz NB Beacon

Power O/p 2-4



Section 8 24GHz

8.1 The Plessey GDO33 Gunn Oscillator

8.1.1 A Power Supply/Modulator for the GDO33 24GHz Oscillator

A power supply/modulator for the GDO33 24GHz oscillator

An increasing number of stations are becoming active on the 24GHz band, most of them using the Plessey GDO33 Gunn oscillator. This oscillator requires a supply of approximately −5V at 250mA, in contrast to most low-power 10GHz oscillators in use which require a positive supply at a somewhat higher voltage. Thus a different power supply is required for the GDO33. G4CNV has supplied details of a suitable design, which is in current use by a number of stations.

The circuit is shown in Fig 1, and one of its features is the inclusion of a current limiter, which makes the power supply short-circuit proof. Tone modulation is provided by a cmos oscillator followed by a simple RC filter. The microphone amplifier was designed for use with a crystal microphone, but should be suitable for most other types. In the unlikely event of more gain being required, the 330Ω resistor in the emitter of the microphone amplifier should be bypassed with a 10μF capacitor.

AFC inputs of either sense are provided, but if a negative earth i.f. strip is used, some form of level translator will be necessary. No connections are made to these inputs if the afc facility is not required.

A pcb layout for the power supply/modulator is shown in Fig 2, and the component layout in Fig 3. Double-sided copper-clad board should be used, the top surface being left copper covered to act as an earthing plane. A number of components are mounted externally, these being the tone/audio selector switch, the TIP2955 transistor (mounted on a heatsink, such as the diecast box used to house the power supply), the microphone gain potentiometer and the output voltage setting potentiometer. The latter should be of the 10-turn variety for ease of tuning, since the available range of output voltage (3·5-5·5V) will tune the oscillator over almost 100MHz.

If the rest of the equipment has a negative earth, it will be necessary to run the GDO33 power/modulator from a separate source, such as another battery. A convenient alternative would be to use a small inverter (with a very well-smoothed output), running from the main +12V supply, to produce the −12V. Any suitable designs for this would be most welcome.

extracted from RadCom

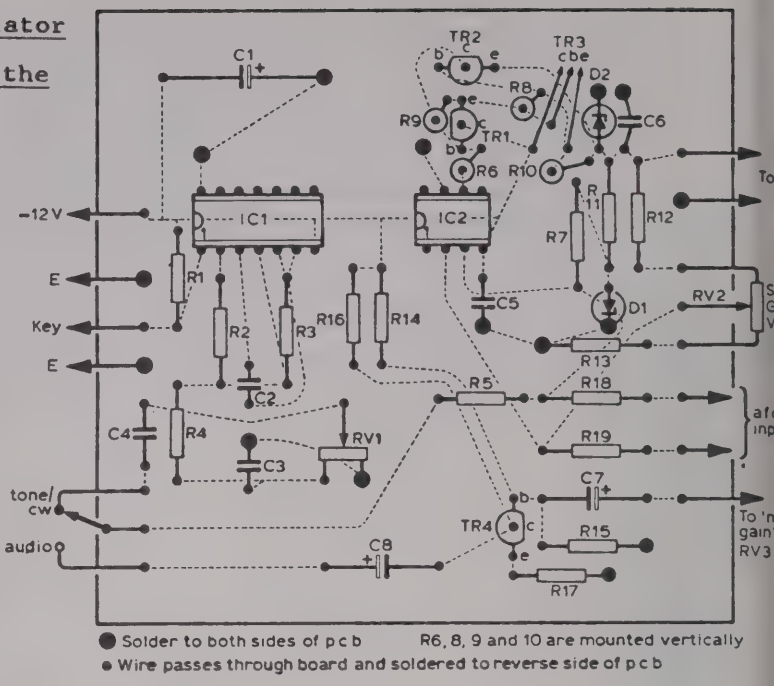


Fig 3. Component layout for GDO33 psu

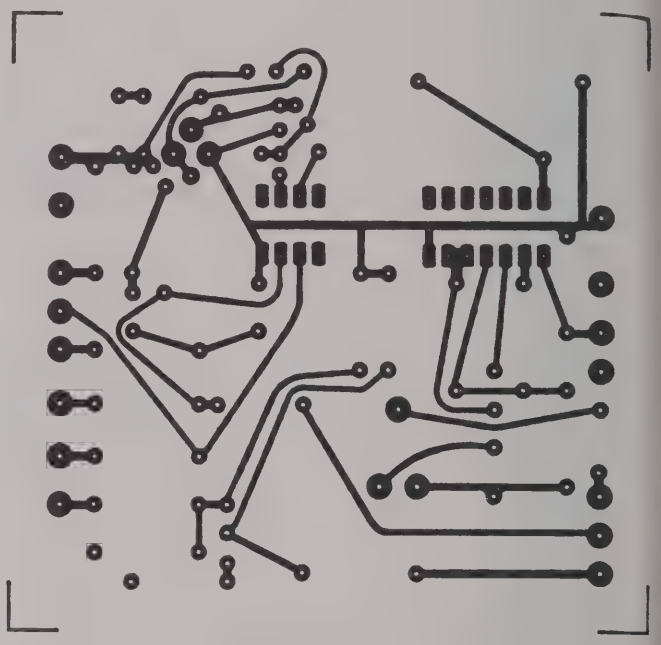


Fig 2. PCB layout for GDO33 psu (copper track side)

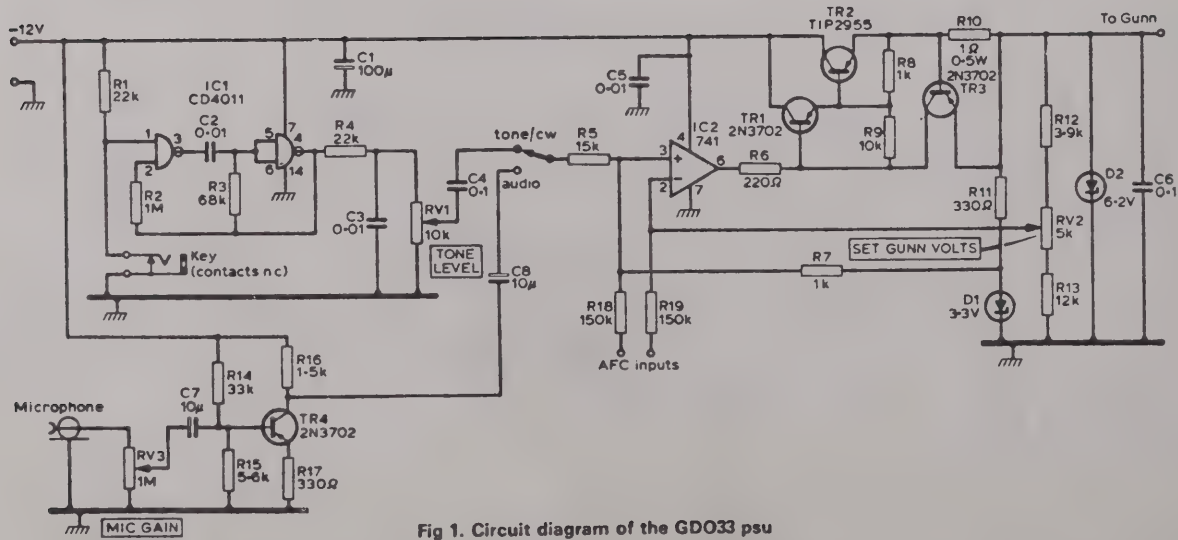


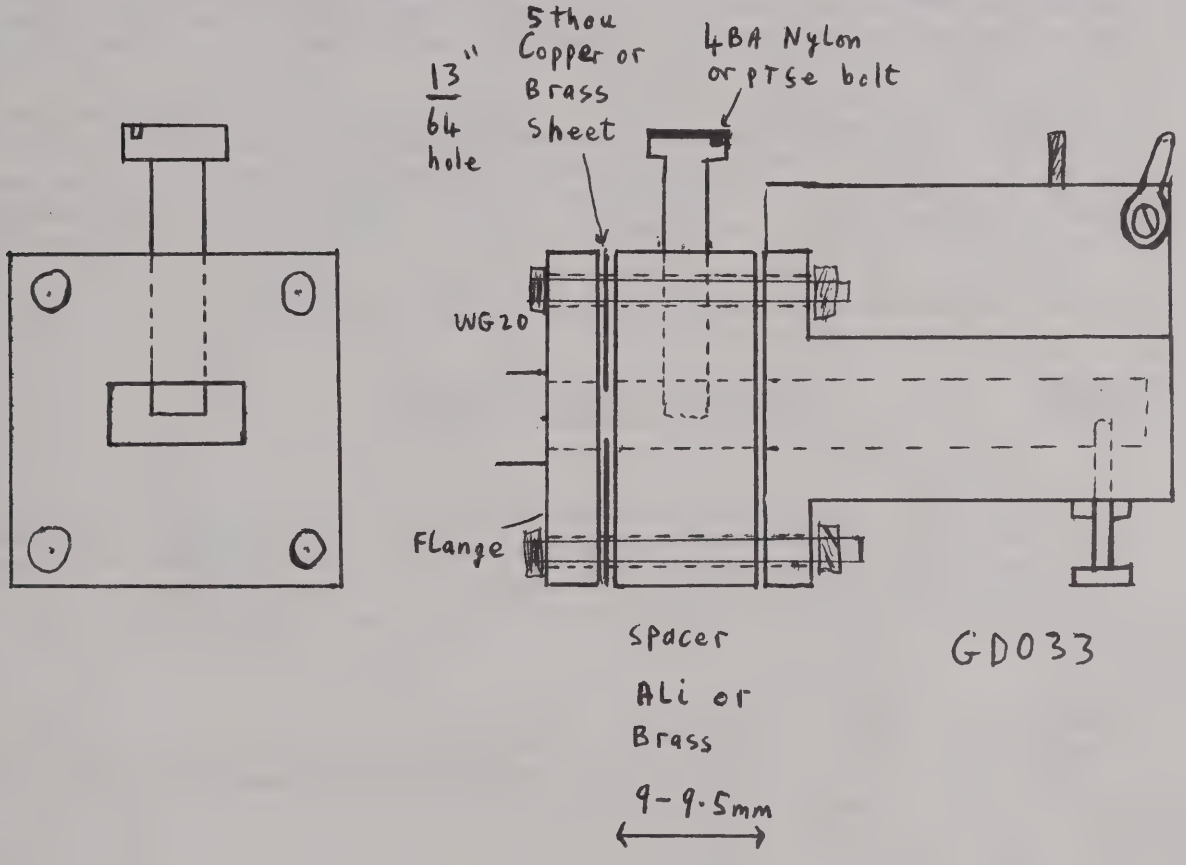
Fig 1. Circuit diagram of the GDO33 psu

Modifications to the GDO33 - G3YGF

Having discovered that the GDO33 could be pulled over 100MHz by a varying load impedance, G3YGF came up with the following design to improve the stability by defining the Gunn cavity with an iris plate.

It is based on the G8APP design for 10GHz, which has a half wave cavity. This would have involved machining the GDO33, and so to avoid this the cavity was extended to 4 lambda, as this was the first multiple of a half wave greater than the length of the GDO33 body where stable operation could be obtained. The modification consists of extending the length of the GDO33 by 9mm using a piece of 9 to 9.5mm brass or aluminium plate as a spacer.

The plate is cut to the size of a flange, and a flange used as a template to drill 4 holes through the spacer. Then several holes are drilled in the center of it, and filed out to WG20 inside dimensions (approx 9 x 4mm). Drill and tap a 4BA (or possibly 2BA) hole in the center of the broad face of the spacer for a tuning screw. The iris plate is made from 3-6 thou brass or copper sheet with a 13/64" hole in the middle of it. All the components are held together by the four bolts through the flanges clamping them to the output waveguide.



Results

1) With the hole diameter shown, the output power was reduced by about 3dB. The hole size could be varied to get either better stability or more output.

2) The tuning range is 200MHz with a 4BA nylon screw; a 2BA screw would give a greater range. The range with PTFE is about half that of nylon.

3) With the 8BA metal tuning screw removed completely, the maximum frequency is about 24.6GHz; it can be used for coarse tuning down to well below 24GHz if required. The range with the nylon screw should be adaquate for most tuning purposes. Voltage can also vary the frequency by about 100 - 200MHz, but this should only be used for very fine tuning as the modulation sensitivity varies greatly with voltage. Once the deviation is set up, the voltage should only be changed by a few hundred mV.

4) The change in frequency as the phase of the load varies around a 1.2:1 vswr circle is reduced from 60MHz to 2MHz.
5) The stability is such that a 25kHz bandwidth receiver has been used for contacts; it drifted out of the bandwidth in about half a minute with the GD033 in open air; a polystyrene box round the GD033 might improve the stabilty further.
6) Any movement of the aerial or of objects near it will affect the frequency, so if you want to work in 25kHz bandwidth be careful to avoid this.
7) The GD033 requires a negative supply voltage, which can be inconvenient. One way round this is to use an existing 10GHz Gunn supply (provided that it is short circuit protected!) and to insulate the GD033 body from earth by placing a thin sheet of PTFE or polythene between the GD033 and the spacer, or between the iris plate and the output flange, and using nylon screws to hold the assembly together. Take care not to apply too high a voltage if using an ordinary 10GHz psu, and try to avoid letting the GD033 body touch any earthed metalwork. Wrapping it in insulating tape may help!
In the last cumulative G4KNZ and G3YGF had a QSO on 24GHz with these mods, using 25kHz bandwidths; although it was only over 4km, it did demonstrate the improved stability.

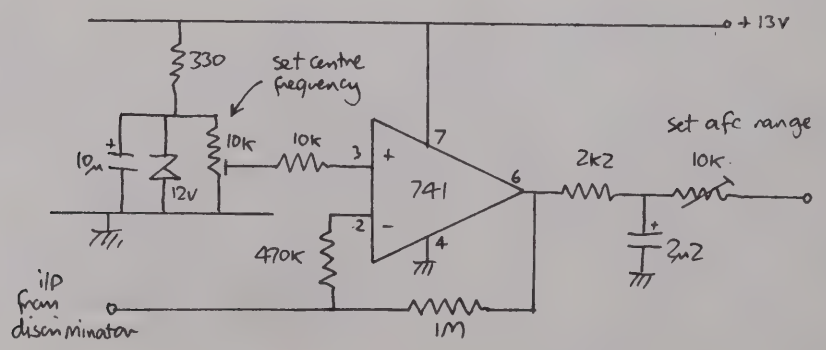
Several people have built the 10dB cross coupler for 24GHz and found that the directivity is very poor. The coupling does seem to be fairly accurate however, so they are still useful for feeding power into the mixer in a receiver. A more directive design is still needed for matching feeds etc.

G4CNV has incorporated G3YGF's iris modification (newsletter 05-81) to his GD033 24GHz oscillator, with a marked increase in stability. He notes that the 9.5mm section can very easily be made from two square flanges with WG20 down the middle. This is exactly the right length. The decrease in power is quite significant, but is more than compensated by the improved stability.

8.1.3 Notes on GD033 PSU

Cliff, G8SHF reports that he found the GD033 24GHz oscillator very sensitive to mismatch. He also had instability problems with the GD033 psu/modulator design (though using different transistors to those specified), but cured it by fitting a 0.1µF capacitor between pins 2 and 6 of the 741.

GW8NBK sends details of the afc system he uses in conjunction with the GD033 24GHz Gunn oscillator power supply/modulator ("Microwaves", Radcom September 1980).



GW8NBK afc system.

8.2 24GHz Components

8.2.1 Diode Mounts

8.2.1.1 An Improved 24GHz Diode Mount - G4CNV

The 1N26 diode mount described in 'Microwaves', February 1980 is a useful basic design, but suffers from a rather inelegant way of making the if/dc connection. This modification avoids the need to connect directly to the diode body; rather than decoupling between the diode body and the diode holder (OBA nut), the two ends of the T-transition are decoupled, and the if/dc connection is taken from one end of the T-crossbar.

The construction should be evident from the diagrams. In all other respects the design is the same as the original.

Two such mounts have been built, one of which is intended for use in the GB3IOW 24GHz beacon, and the whole mount including matching screws is just 30mm (1.2 inches) long.

If the unit is used as a mixer rather than a detector, it is good practice to build the if preamplifier stage directly onto the mixer mount, so that the matching between the mixer and preamplifier is constant once optimised, and not affected by random lengths of cable that might otherwise be used to connect mixer and preamp.

One standard design of if preamplifier is the untuned 2 x BF180 circuit (VHF/UHF Manual, 3rd edition, p8.39), which is usable at the common if's of 10.7, 30, 100 or 144MHz. This design has been quite widely used, mostly at 10GHz, where the ability to cope with if's of different frequencies has sometimes been useful, when, for instance, a narrow-band 144MHz receiver could easily be substituted for the usual wideband fm if strip.

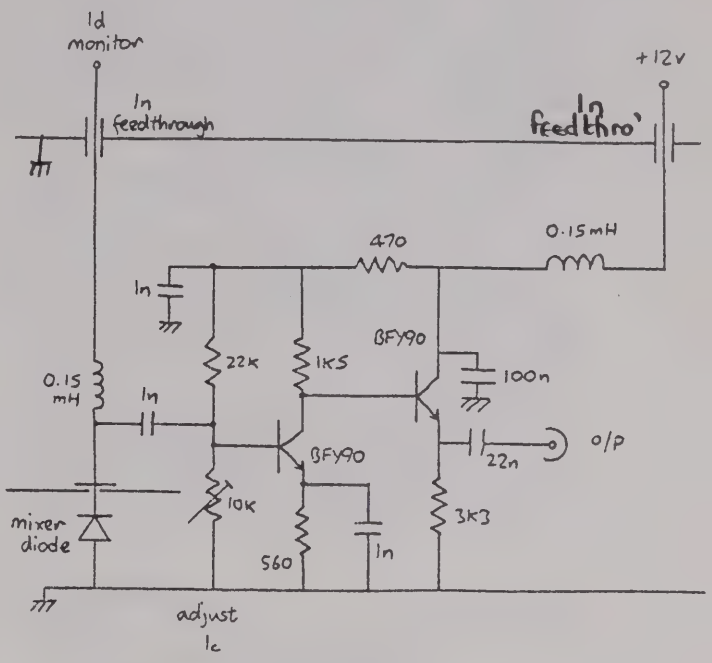
The basic design does not appear to be optimum, and it seems worthwhile to incorporate some means of adjusting the Ic of the first transistor to optimise the receiver performance.

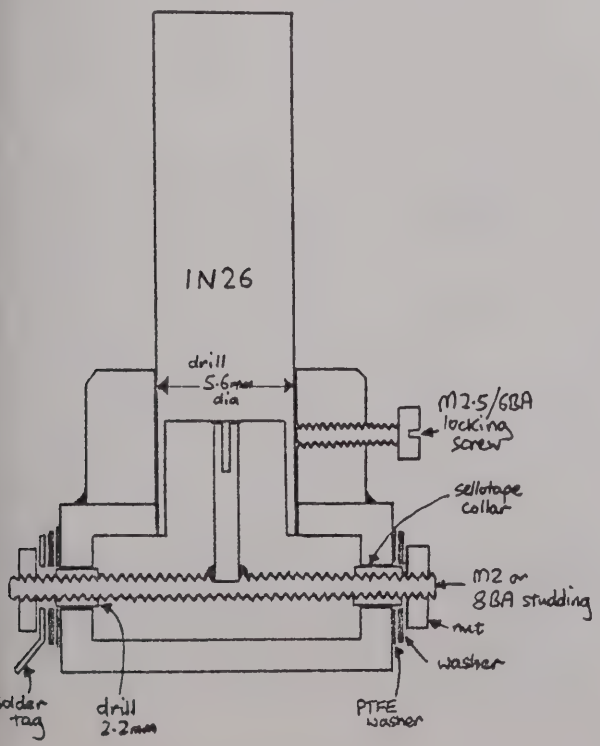
Adjusting the Ic in this way has two effects - firstly the transistor noise figure will occur at a particular value of Ic, and secondly the transistor input impedance will vary with Ic and can therefore be adjusted to provide the best match to the diode, whose video impedance will be of the order of a few hundred ohms.

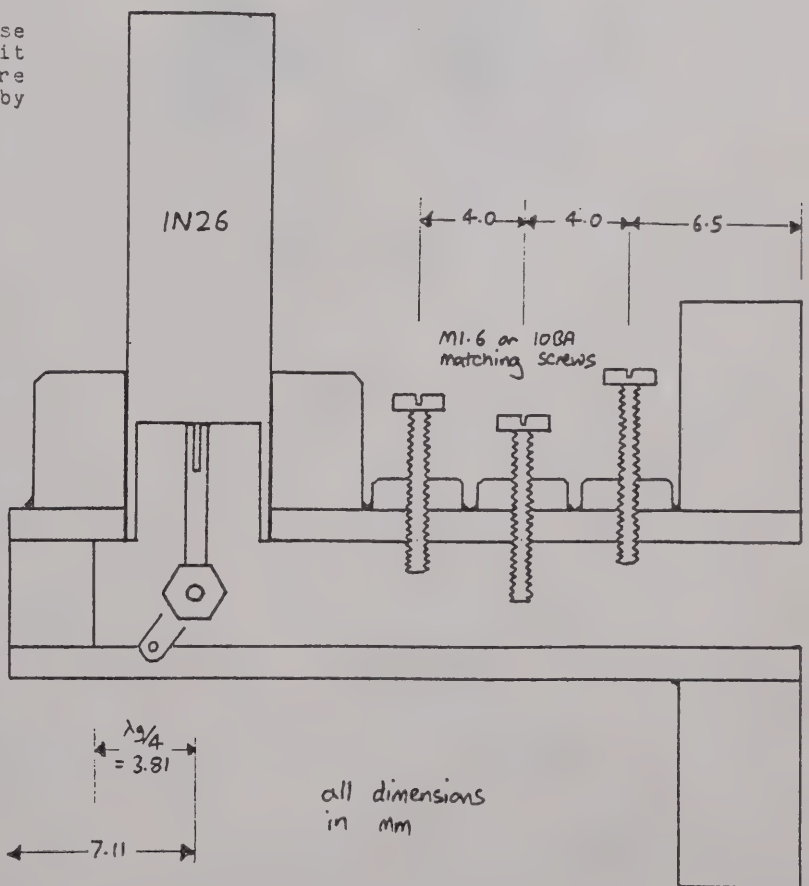
The adjustment will, of course, be a compromise between these two effects, and the mixer unit should ideally be aligned using a noise figure indicator, though it is possible to optimise by ear using a weak signal.

There are several variables in the mixer assembly that can be tweaked; lo level, rf matching (screws), mixer-preamplifier matching, preamplifier noise figure, and if several mixer diodes are available it may be possible to select the best of the batch. G3WDG has demonstrated with 10GHz wideband receivers that there are usually many dB's of performance to be won by careful optimisation. The potential improvement with 24GHz receivers is probably even greater, though it has not yet been possible to use a IN26 diode as a noise source in the same way as at 10GHz.

The improved preamplifier design, which uses BFY90 transistors, is given in the diagram below.







8.2.1.2

A 1N26 mixer for 24GHz

This mixer is constructed out of WG20 and can be built using a vertical bench drill and hand tools only, no lathe work being necessary. Constructional details are shown in Fig 1. The writer's prototype measures only 1·5in long, including a home-made flange.

First, square off the end of the waveguide, by marking out with a right angle, and filing to the marks. A 0BA hole is then drilled and tapped 0·25in from the squared-off end. The tapping should be commenced using a tapered tap, changing over to a plug tap when the tapered tap comes up against the opposite wall. Assemble a 0BA nut on a 0BA screw, the latter preferably cadmium plated or sufficiently dirty so as not to take solder readily. Screw this a short distance into the waveguide and tighten up the nut against the waveguide wall. Using a hot-plate or small flame to provide heat, solder the nut in place. When cool, remove the screw and drill out the nut and waveguide wall underneath the nut using a 5·7mm drill. Drill and tap the nut 6BA through one of its faces, and remove the burr on the inside using the 5·7mm drill.

Drill with an 8BA tapping drill through the centre of the narrow wall of the waveguide, 0·25in from the end, continuing through the opposite wall. Open up one of the holes to 8BA clearance, and tap the other 8BA. Assemble a 0·5in 8BA brass screw through the clearance hole and screw tightly into the opposite wall. Using the 5·7mm drill inserted through the 0BA nut, make a mark on the 8BA screw. Remove the screw and, using a small needle file, make a 0·090in-wide slot in the screw, centred on the marked point, to half-way through its diameter.

Next, prepare the centre conductor for the 1N26 in the following way. Remove the centre connecting piece from a cable-mounting bnc socket, saw off the end remote from the metal fingers, and file to length (0·235in). *Carefully* squeeze the fingers inwards at a point roughly half-way down their length, using the wire-cutting edges of a pair of pliers, so that an opposite pair of fingers are nearly touching. The connecting piece should then be a good push fit on the centre pin of the 1N26. An alternative connecting piece which does not require such modification can be one of the contacts from the high-quality ptfe type of transistor holder.

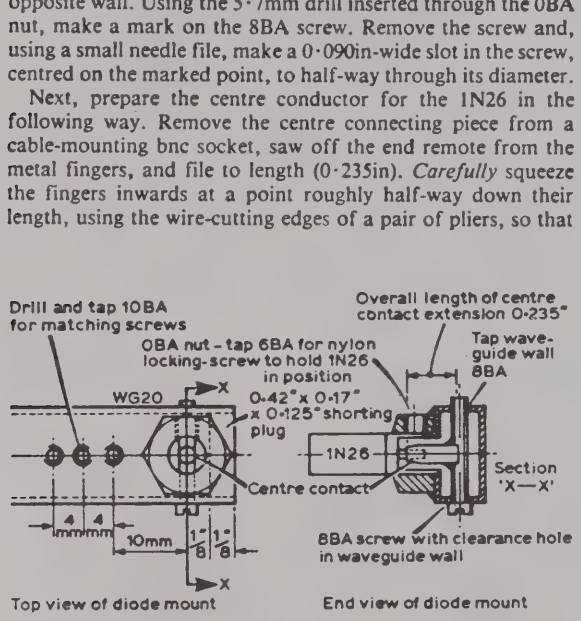


Fig 1. A 1N26 mixer for 24GHz

Wrap one layer of Sellotape around the bottom half of the 1N26, to insulate it when it is in operation, and push the modified centre connector onto the pin of the diode, making sure that it is fully home. Reinsert the 8BA screw into the waveguide, tightening until the filed slot is uppermost. Insert the 1N26 diode into the 0BA nut, pushing in until the centre connector engages into the filed slot in the 8BA screw. The end of the outer casing of the diode should then be approximately level with the inside wall of the waveguide. Using as large a soldering iron as will fit, quickly solder the centre contact to the 8BA screw, using a minimum of solder. Ensure that not too large a blob of solder remains on the joint. As soon as the solder solidifies, remove the diode to reduce the chance of damage, and check that the layer of Sellotape has not been damaged.

The shorting plug is then made by sawing out a piece of 0·125in brass sheet and carefully filing to size. Constant checking of the dimensions with a micrometer during filing will ensure accuracy. The last few thousandths of an inch are best removed by rubbing the plug on emery-paper, or wet and dry paper, laid on a flat surface. The plug is then pressed into the end of the waveguide with the aid of a vice. Finally, drill and tap the 10BA holes for the matching screws, and remove burrs with a file.

The setting-up of the unit is straightforward; the matching screws simply being adjusted for maximum diode current (measured between the body of the diode and the waveguide). The screws can be retained in their optimum positions with lock-nuts. Permanent connection to the body of the diode can be made by soldering a wire to the top of the body, but a safer method would be to use a spring clip, or a small collar fitted with a locking screw, to which a wire could be soldered beforehand. This would avoid the risk of overheating the diode.

extracted from RadCom

Control of a Gunn Oscillator by Phase Locking Techniques

G3BNL

Introduction

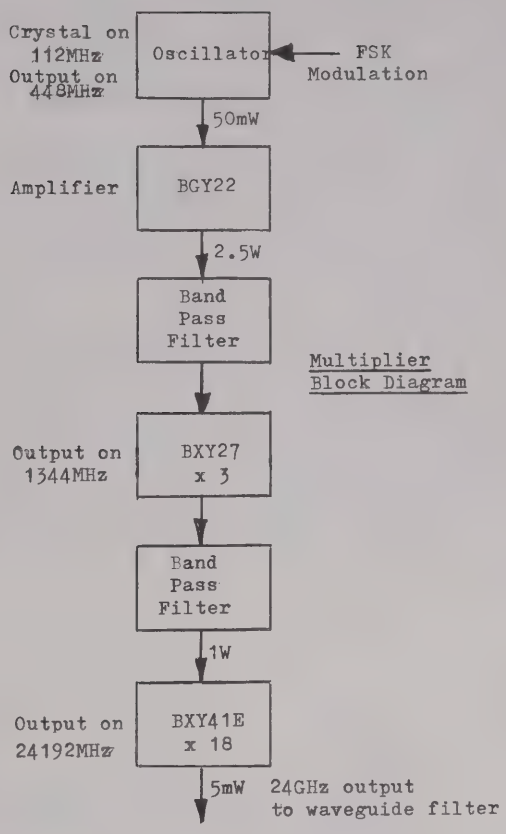
Operation on the higher microwave bands falls into two basic catagories:

1. Narrow band operation with a crystal controlled oscillator.
2. Wide band operation using a free running oscillator.

Both methods have their followers and we may comment as follows:

1. Narrow band operation
   a. The frequency may be set and known to narrow limits removing one of the uncertanties of portable operation.
   b. Drift can be minimal and tolerable.
   c. System performance is enhanced due to the narrow bandwidth which can be used for receiving.
   d. To achieve a high output power (eg 100mW) it is necessary to start with a lot of power at 384MHz.
   e. Varactors in cascade can be difficult to maintain in a stable condition without test equipment.
2. Wide band operation
   a. Gunn oscillators are available for powers of up to ½W.
   b. Drift is troublesome and frequency stability gets worse for higher powers.
   c. Wide band receivers have to be used with an inferior performance to CW or SSB receivers.
   d. This type of equipment is easier to construct and get operational.

Thus it can be seen that both methods have advantages and disadvantages. What is proposed is to combine the better features of both methods by phase locking a Gunn diode oscillator to a low power crystal controlled source. By this means the Gunn device takes on the stability of the crystal source while providing a higher output power than would be obtained from the crystal source on its own. With 2 to 3mW from the crystal source it is possible to control a Gunn diode with a power output of up to ½W. Also, once the phase locking circuitry is constructed it may be used on any microwave band for which a Gunn diode and low power crystal controlled source is available.



24GHz System

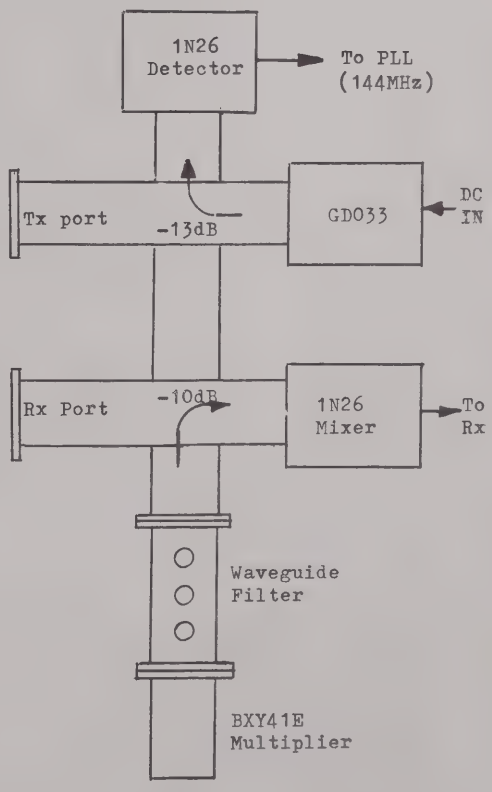
From the block diagram for the 24GHz NB system it can be seen that the low power crystal controlled source starts off with a crystal at 112.000MHz. This is used in a board similar to the UHF source described in October 81 Radcom. It is useful to add a varicap diode to the oscillator to provide FSK modulation facilities.

The output at 448.000MHz is fed to a BGY22 amplifier mounted on a small heatsink (via an attenuator as the device only requires 50mW drive). The output from the amplifier is taken via a bandpass filter to a varactor multiplier (x3) and to a further bandpass filter. The output at this point is about 1W at 1344MHz. A further multiplier follows which multiplies to 24192MHz in one step. This multiplier is constructed in WG20 and is followed by a waveguide filter which is a scaled down version of the 10GHz JVL type. The power output is about 3mW.

For the receiver there is more than enough power for driving a low noise mixer such as a 1N26 and a directional coupler is used (6 to 10dB) to couple power into the mixer. The local oscillator of 24192MHz is intended to provide reception on 24046 to 24048MHz when used with a tuneable IF of 144 to 146MHz.

On transmit the remainder of the power from the low power source is fed into another 1N26 mixer together with a portion of the output from the Gunn diode oscillator fed from a directional coupler. The Gunn is set roughly on the transmit frequency of 24048MHz. The signals in this second mixer thus produce an output of roughly 144MHz. This signal is then fed to the PLL circuitry for locking. When lock is established the Gunn frequency will be maintained exactly 144MHz above the frequency of the low power source ie exactly on 24048MHz and any drift in the Gunn will be automatically corrected for.

[Note that a working frequency of 24048MHz is being used. This is because the low power crystal controlled source is locked to a frequency standard. It is possible to implement this type of system for a working frequency of 24192MHz by using a crystal controlled source on 24048MHz. However this is not a convenient multiple of 200kHz. In this case the crystal required would be 111.333MHz.]



Block diagram of waveguide components

Phase Lock Unit

A low noise crystal oscillator circuit at 1MHz is implemented using a MC12060. The output from this IC needs to be brought up to the CMOS input level of the phase comparator MC14568 and a BC108 is used for this purpose.

The MC14568 also has two counters within the chip and the 1MHz is first divided down to 250kHz before the phase comparison.

Looking at the input from the 1N26 mixer the low level signal is first fed to a low noise preamplifier to give enough input signal for the SP8659 divider chip. This is a 200MHz divide by 16 ECL package with an open collector output. This is followed by a TTL divide by 4 stage using a 74LS113 thus giving a frequency of 2.25MHz. The output is then shifted in level using a BC108 to produce CMOS voltage levels.

The second counter in the MC14568 is then used to divide the signal by 9 to give a frequency of 250kHz also.

The output of the phase comparitor estabishes a voltage across a suitable integrator which is used to control the voltage applied to the Gunn diode. Following the integrator (charge pump) it is necessary to filter out the 250kHz pulses present and this is done with a twin "T" Notch filter followed by a low pass filter. After this loop filter there is a DC buffer formed from a 741 op amp.

It is important for the output to be thoroughly filtered as any noise will result in the Gunn being frequency modulated.

A further 741 is used in conjunction with a 2N3055 to provide the power supply to the Gunn diode. The 2N3055 is operated as an emitter follower with provision to adjust the initial setting of the supply output to bring the diode approximately onto frequency before locking. It is wise to prevent the output voltage rising beyond the maximum rating of the Gunn diode by the use of a zener on the input to the 2N3055.

Modulation is applied to the diode supply from either a microphone preamplifier or a 1kHz tone oscillator. The deviation required for narrow band FM is about ± 3kHz.

In practise it is necessary to set the Gunn diode to within about 30MHz of the required frequency by either mechanical tuning or adjustment of its supply. Once this has been done once, it should only be necessary to power up the PLL circuitry and lock should be established without need for further adjustments.

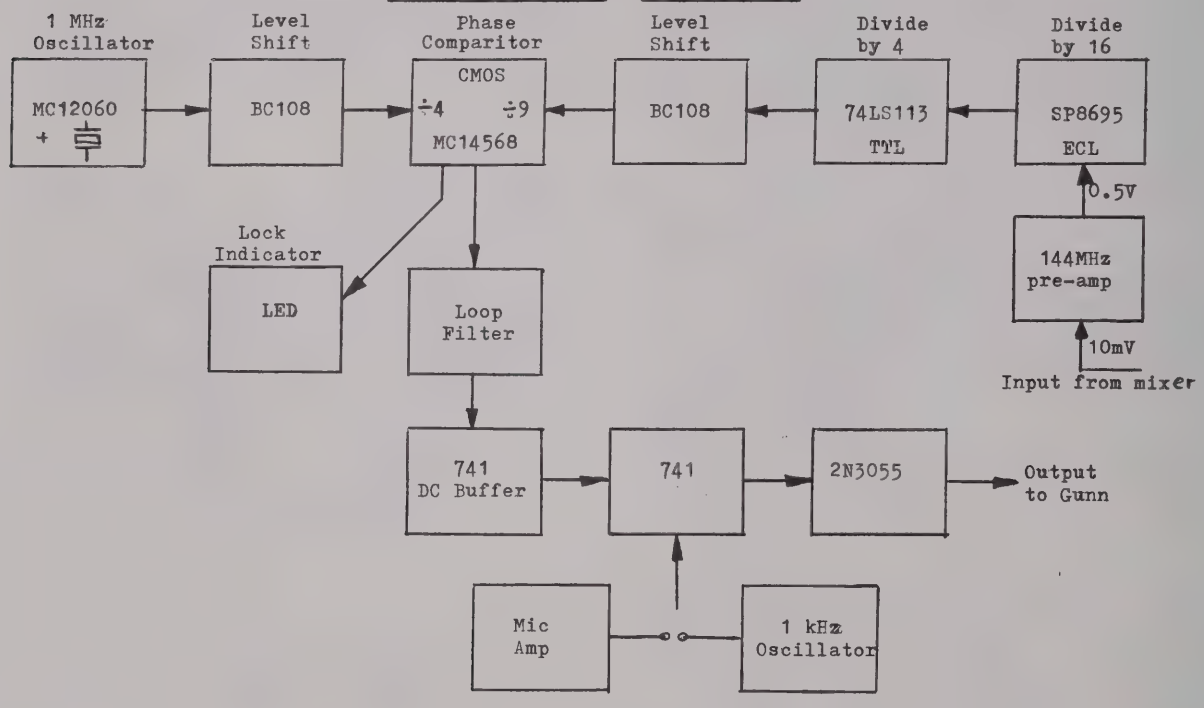
A meter can be provided to monitor the Gunn supply voltage and the fact that the system is locked can be indicated by an LED driven by the MC14568 phase comparitor which is wired to extinguish when locked. Further proof that the system is in lock can be obtained by tuning the receiver IF down to 144MHz exactly when the Gunn should be clearly audible.

Conclusion

The results so far have been promising. As stated earlier the method can be applied to any of the microwave bands and is a very cost effective means of getting high power narrow band equipment going.

Further details can be obtained from the author, G3BNL who is QTHR.

Phase Locking Unit - Block Diagram



Section 9 Millimetric Bands

9.1 Introduction — Millimetric Bands - G4KNZ

At WARC 79, amateurs were allocated five new millimetric bands - 47, 76, 120, 142 and 241 GHz. Although there is much still to be learnt on 24GHz and 10GHz, and indeed lower microwave bands, it would be a very worthwhile achievement simply to make a qso on these new bands. Very little information is available on designs for equipment and most of the initial work will be experimental. Here is a summary of some data and a few ideas for 47GHz and 76GHz in the hope that it will encourage a few readers into action building gear. Particular thanks are due to G3HWR for the absorption data.

1. WARC Allocations

47.0 to 47.2GHz exclusive

75.5 to 76.0GHz exclusive, 76.0 to 81.0GHz secondary

No news of availability in the UK yet.

2. Frequency of Operation

For crystal controlled equipment, a frequency of 76032MHz suggests itself, being 1152 times 66. No useful relation has been found for the 47GHz band (the nearest multiple of 1152 is just outside the band at 47232MHz) and the only suggestion is to use a multiple of 96MHz ie 47040 or 47136MHz.

A free running oscillator (ie a Gunn) on 47GHz would easily tune the whole band. On 76GHz, this probably would not be the case and so it might seem sensible to aim to operate fairly close to 76032MHz.

9.2 Propagation

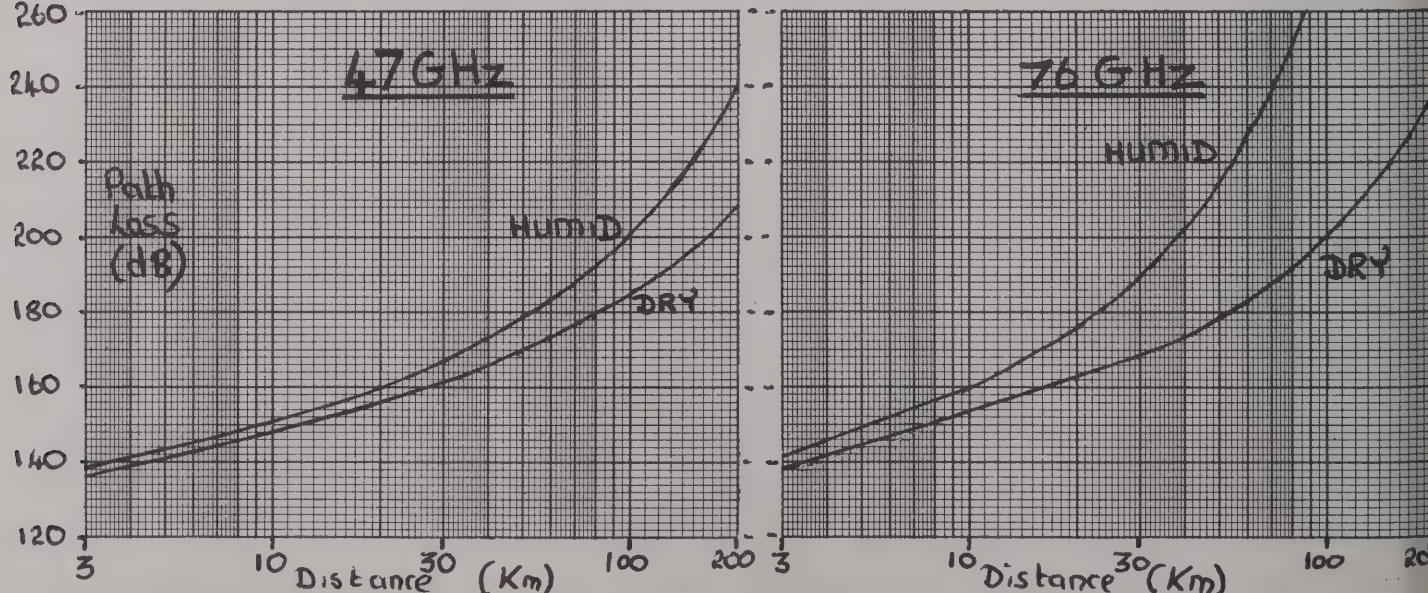
9.2.1 Atmospheric Attenuation

On the centimetric bands, any optical path in the UK can be worked with even simple equipment. This would also be true on 47GHz and 76GHz if it were not for absorption in the atmosphere. In dry air, absorption is mainly due to oxygen, but significant additional attenuation is caused by water vapour. What matters here is the absolute mass of water vapour present in the atmosphere so that attenuation may be more severe on a humid summer day than a wet but cold winter day. If the temperature is below freezing the water content of the atmosphere is very low - this might be the best time to operate! Some approximate figures for the attenuation and the comparative ones for 10GHz and 24GHz are shown below. These do not include rain across the path, which will cause the attenuation to increase very considerably at 47GHz and above.

Band (GHz)	Attenuation (dB/km)		
	Dry Air	Moderate Humidity	High Humidity
10	0.015	0.02	0.03
24	0.022	0.16	0.30
47	0.18	0.20	0.35
76	0.3	0.4	1.0

9.2.2 Path Losses

These figures can be combined with the free space loss to obtain approximate path loss figures as a function of distance. These are plotted on the two graphs below for 47GHz and 76GHz. The loss figures represent the loss between two isotropic antennas over the corresponding distance. Note that the humid air figures are for high humidity.



9.3 Ideas for Getting Started

The simplest suggestion is to use a mixer diode as both a mixer (on receive) and a multiplier (on transmit and receive) where the diode acts as a doubler or tripler. Thus on 47GHz a signal at 23.5GHz from either a Gunn or varactor diode could be doubled. This might be implemented using some sort of coupler arrangement, made from two different waveguide sizes (eg WG20 & WG24) with the mixer diode sitting in the junction. The IF could be taken from the diode to a conventional IF strip.

G3WDG points out that the output from a Gunn diode usually contains a considerable amount of the 2nd harmonic - it may only be 10 or 15dB down on the fundamental. Thus a GD033 module could be retuned to 23.5GHz and a simple filter placed on the output in WG24 to reject the fundamental. Thus a low power tx or local oscillator is easily built.

For an antenna gain of 37dB, which corresponds to a 3' dish on 10GHz, the dish sizes on 47GHz and 76GHz are 12 and 7 inches approximately. However, horns are quite practical antennas at these frequencies and are especially attractive as no critical feed matching is needed. The aperture sizes for a gain of 30dBi are 9cm by 7.5cm on 47GHz and 5.5cm by 4.5cm on 76GHz - similar in size to a 15dBi horn on 10GHz. These sizes will give a beamwidth of about 5 degrees, corresponding to a dish size of 18 inches on 10GHz.

The standard waveguides are listed in the VHF/UHF manual, but an alternative are the brass rectangular sections mentioned by G3HWR

G3HWR has had some interesting ideas about waveguide for the new millimetre bands :-

> "Waveguide sizes are agreed internationally for frequencies up to at least 300GHz and for amateurs with access to them, these are the best to use.
>
> For home built equipment which does not require to connect to other equipment, it is not necessary to use standard waveguide and non standard waveguide quite adequate in short lengths can be bent up from sheet metal.
>
> Useful extruded metal sections are available from the model trade, these are used by modellers to represent the corresponding steel sections ie., angle, I-beam, T-beam and square and rectangular box sections!

These extruded sections used by the model trade are available from R. M. Trading Co., 646 High Road, N Finchley, London N12 ONL, Telephone 01-445-6531.

> Attached is a list of the important properties of the four rectangular sections and the waveguide obtained when a channel section is attached to a flat sheet."

For getting going on the millimetric bands, G3JHM suggests the use of sub-harmonic mixing: $F_{sig} - 2x\ F_{lo} = F_{if}$. This will enable existing multipliers to 24GHz to be used for a 48GHz downconverter, for example. The sub harmonic mixing enables a simpler design of mixer circuit to be used since the frequency separation between F_{sig} and F_{lo} is much greater. This technique has been studied by several workers at BTL and TRG (Alpha) and may be applied from 1GHz to at least 183GHz. Don says he has a bibliography and copies of the relevant papers if anyone is interested. In general, the realisation of the circuits is by using stripline mounted symetrically in waveguide, with the base material being quartz substrate but it may be possible to use polyguide or other microwave pcb materials.

9.4. Metal Sections for use as Waveguides - G3HWR

Properties of Non Standard Waveguide

Cat No K&S	Description Type	O.D. in. mm	I.D. in. mm	Cut Off GHz	Useful Band (GHz)	Aerial Band GHz	λ_{air} (mm)	λ_g	λ_g/λ_a
262	□	.187 x .093 4.76 x 2.38	.156 x .063 3.96 x 1.60	37.8	40 to 75	47 47.2			
264	□	.250 x .125 6.35 x 3.18	.219 x .094 5.55 x 2.38	26.98	30 to 52	47 47.2	.6378 .635	.7785 .7734	1.221 1.217
266	□	.313 x .156 7.93 x 3.97	.281 x .125 7.14 x 3.18	21.0	24 to 41	24.0 24.192 24.25	1.249 1.239 1.236	25.72 24.96 24.65	2.060 2.014 1.995
268	□	.375 x .188 9.53 x 4.76	.344 x .156 8.74 x 4.0	17.17	20 to 34	24.0 24.192 24.25	1.249 1.239 1.236	18.18 17.86 17.78	1.454 1.442 1.438
181	⊓	.125 x 3.175 x	.0937 x 2.38 x	62.94	70 to 125	76. 76.03 81 120	3.94 3.70 2.50	7.029 5.878 2.936	1.784 1.588 1.174
184	⊓	.219 x 5.56	.188 x 4.76	31.47	35 to 60	47 47.2	6.38 6.35	8.588 8.522	1.346 1.342

K&S METALS AND TOOLS

Send to: R.M. Trading Co., 646 High Road, N. Finchley, London N12 0NL, Tel: 01-445 6531

EXCLUSIVE U.K. IMPORTERS OF K & S PRECISION ENGINEERING PRODUCTS

Metal Centre

Complete kit, inc. display stand contains a selection of the following items.

ALUMINIUM TUBE (12 Inch Lengths)

Ref No.	Quantity	No. In Pack	Size	Price Per Item	Total
100		15	1/16	.25	
101		12	3/32	.30	
102		12	1/8	.30	
103		12	5/32	.35	
104		12	3/16	.40	
106		10	7/32	.45	
106		10	1/4	.50	
107		8	9/32	.65	

COPPER TUBE (12 Inch Lengths)

*117		20	1/16	.25	
*118		15	3/32	.30	
*119		12	5/32	.40	
120		12	1/8	.30	

SOFT BRASS FUEL TUBE (12 Inch Lengths)

121		10	1/8	.40	

BRASS STREAMLINE (12 Inch Lengths)

122		4	Small	.75	

ROUND BRASS TUBE (12 Inch Lengths)

125		20	1/16	.30	
126		15	3/32	.30	
127		15	1/8	.30	
128		12	5/32	.35	
129		12	3/16	.45	
130		10	7/32	.50	
131		8	1/4	.55	
132		8	9/32	.60	
133		6	5/16	.65	
134		6	11/32	.70	
135		6	3/8	.75	
136		4	13/32	.85	
137		4	7/16	.90	
138		4	15/32	.95	
139		4	1/2	1.00	
140		3	17/32	1.05	
141		3	9/16	1.10	
142		2	19/32	1.20	
143		2	5/8	1.25	
144		2	21/32	1.40	

SQUARE BRASS TUBE (12 Inch Lengths)

149		12	1/16 sq.	.45	
150		12	3/32 sq.	.50	
151		12	1/8 sq.	.55	
152		10	5/32 sq.	.65	
153		8	3/16 sq.	.75	
154		6	7/32 sq.	.85	
155		6	1/4 sq.	.95	

SOLID BRASS ROD (12 Inch Lengths)

159		50	.020	.08	
160		50	1/32	.08	
161		34	3/64	.12	
162		20	1/16	.20	
163		16	3/32	.25	
164		10	1/8	.40	
166		8	5/32	.50	

BRASS ANGLE (12 Inch Lengths)

171		10	1/8 x 1/8	.40	
172		10	5/32 x 5/32	.45	
173		8	3/16 x 3/16	.50	
174		7	7/32 x 7/32	.55	
175		6	1/4 x 1/4	.65	

BRASS CHANNEL (12 Inch Lengths)

181		10	1/8	.50	
182		8	5/32	.55	
183		7	3/16	.60	
184		6	7/32	.65	
185		5	1/4	.75	

Total

ROUND PLATED SPRING WIRE (12 Inch)

Ref No	Quantity	No. In Pack	Size	Price Per Item	Total
192		50	.032	.08	
195		50	047	.08	
197		50	055	.08	
199		50	063	.08	

BRASS STRIPS (12 Inch Lengths)

230		20	016 x 1/4	.20	
231		15	016 x 1/2	.30	
232		10	016 x 1	.50	
233		10	016 x 3/4	.40	
234		5	016 x 2	.90	
235		15	025 x 1/4	.25	
236		10	025 x 1/2	.40	
237		6	025 x 1	.70	
238		6	025 x 3/4	.55	
239		3	025 x 2	1.10	
240		15	032 x 1/4	.30	
241		10	032 x 1/2	.50	
242		5	032 x 1	.85	
243		6	032 x 3/4	.65	
244		3	032 x 2	1.60	
245		8	064 x 1/4	.60	
246		4	064 x 1/2	1.00	
247		3	064 x 3/4	1.25	
248		2	064 x 1	1.70	
249		3	064 x 2	3.00	

SHEET METAL (4 x 10 Sheets)

250		6	005 Brass	.70	
251		6	010 Brass	1.10	
252		6	015 Brass	1.50	
253		3	032 Brass	2.70	
254		6	008 Tin	.50	
255		6	016 Alum	.50	
256		6	032 Alum	.80	
257		6	064 Alum	1.35	
258		6	Asst Brass	1.30	
259		3	.025 Copper	2.60	

RECTANGULAR BRASS TUBE (12 Inch Lengths)

262		4	3/32 x 3/16	1.10	
264		4	1/8 x 1/4	1.20	
266		4	5/32 x 5/16	1.30	
268		4	3/16 x 3/8	1.40	

HEXAGON BRASS TUBE (12 Inch Lengths)

*271		8	3/32	.55	
*272		7	1/8	.65	
*273		6	5/32	.75	
*274		5	3/16	.85	

Total Carried Forward

Total

Wire & Tube Centre

Complete kit, inc. display stand contains a selection of the following items.

PIANO WIRE (36 Inch Lengths)

Ref No.	Quantity	No. In Pack	Size	Price Per Item	Total
498		50	015	.10	
499		42	020	.12	
500		34	025	.15	
501		34	032 (1/32")	.15	
502		34	047	.15	
503		32	055	.18	
504		30	062 (1/16")	.17	
505		25	078	.20	
506		17	3/32"	.30	
507		12	1/8"	.45	
508		8	5/32"	.65	
509		5	3/16"	1.00	
† 510		4	7/32"	1.25	
† 511		3	1/4"	1.70	

STREAMLINE TUBE (35 Inch Lengths)

† 1100		5	1/4	1.50	
1101		5	5/16	1.80	
1102		4	3/8	2.10	
1103		4	1/2	2.40	
† 1104		3	5/8	2.70	
† 1105		2	3/4	3.00	

ROUND ALUMINIUM TUBE (36 Inch Lengths)

† 1108		8	3/32	.90	
1109		8	1/8	.90	
1110		6	5/32	1.05	
1111		6	3/16	1.20	
1112		6	7/32	1.35	
† 1113		5	1/4	1.50	
† 1114		5	9/32	1.65	
† 1115		5	5/16	1.80	

POUND BRASS TUBE (36 Inch Lengths)

1143		10	1/16	.90	
1144		8	3/32	.90	
1145		8	1/8	.90	
1146		6	5/32	1.05	
1147		6	3/16	1.35	
† 1148		5	7/32	1.50	
† 1149		5	1/4	1.65	

Tools and Accessories

296		3	Tubing Cutter	3.95	
321		6	Tubing Bender Kit	1.50	
702		6	Brass Connectors	1.00	

DISPLAY RACKS

1005			Metal Centre	309.00	
2400			Tube & Wire Centre	169.00	
5002			Easy Rider for 1005	64.00	
(See items with *)					

Total Carried Forward

Sub Total

Retail P + P 35p

Grand Total

†Not included in Wire and Tube Centre.

*included in Easy Rider add on set for 1005 Metal Centre.

In the interest of economy and speed of service, retailers must order in pack quantities only. Retail direct mail orders in any quantity required.

Section 10 Oscillators

10.1 A high-quality uhf source for microwave applications

by THE RSGB MICROWAVE COMMITTEE

10.1.1 page1 Constructional Details

Introduction

This article describes a unit to deliver a minimum of 100mW from a 12V supply, at an output frequency of 360–440MHz, using a fifth overtone crystal in a range 90–110MHz. Its output is intended for multiplication up to microwave frequencies for use as a local oscillator or transmitter, and still has a good quality note at 10GHz. It has been developed by members of the RSGB Microwave Committee from the Plessey AMETS transmitter board.

The design uses a Butler crystal oscillator, which eliminates frequency multiplier stages and gives a very-low-noise output compared with the commonly-used single-transistor circuits. This is due to the loading on the crystal being less, thus allowing a higher working Q to be realized. Leadless disc capacitors are used on a double-sided pcb for efficient decoupling, and the two amplifier stages are run in Class A, contributing to the low-noise performance and making problems with instability very unlikely. All inductors and lines are printed on the board, and only four adjustments are needed to align the unit. Provision is made for on/off keying of the output stage, or for applying fm or fsk to the oscillator. Several dozen of these boards have been built and used in a variety of microwave applications, giving excellent results.

Construction

The unit is constructed on 0·063in (1·59mm) thick double-sided copper-clad glassfibre epoxy board (dielectric constant ≈5) with an earthplane on the component side of the board. The use of materials with different dielectric constants could result in incorrect resonant frequencies for the tuned circuits, and should be avoided. The circuit diagram is given in Fig 1, and the artwork for the two sides of the pcb is shown in Fig 2.

Construction is generally straightforward; the layout of components on the pcb is shown in Fig 3. It is important that the component values are adhered to, and in particular that only new, branded semiconductors are used for TR1–4. Surplus types do not usually perform satisfactorily. The earthed ends of the components are soldered to both top and bottom earthplanes, and all components should be mounted with absolute minimum lead lengths. TR2, TR3 and TR4 should be pushed right down on to the top side of the board, and the cans of TR2 and TR3 soldered to the earthplane. TR4 is fitted with a TO5 heatsink.

Care should be taken in the mounting of the 1,000pF leadless d ceramic capacitors. They are located in slots in the pcb made by drilli several holes close together and then joining them to form a slot. This c be done quite easily using the tool shown in Fig 4, which is made by taki a junior hacksaw blade, breaking off the end with the pin through it, a cutting or filing the end to a taper. The teeth should be oriented as show The point is thin enough to fit into one hole at the end of the row a enable it to be opened up, and then the remaining holes can be joined progressively. The hacksaw blade must be held quite close to the boar and used gently to start with to prevent it snapping off; if it does, it c easily be repaired by filing it to a point again. Another method is to use 1mm drill bit in a vertical drill as a mill. Hold the drill bit in the chuck only 3 or 4mm protrude, and move the board sideways so that the edge the drill bit cuts the slot, joining the holes. Hold the board against straight edge to ensure a straight slot. If the drill breaks, it can still be r used. The edges of the slot can then be cleaned up with a flat needle file nail file. Do not make the slot too wide, otherwise it will be difficult solder the discs, which should be positioned and soldered exactly as show in Fig 5. Ensure that the solder flows properly on to the metallization c the capacitors. Various "surplus" leadless discs gave trouble in th respect on some of the prototypes, and the most reliable types were four to be those made by Steatite[1] which are actually trapezoidal (coffi shaped) rather than round. They proved extremely easy to solder in plac and are thoroughly recommended.

For other types, a useful technique is to heat the edge of the slot first, in sert the disc, apply the soldering iron to the disc (not the board), and then apply solder to form a neat fillet. If silver-loaded solder can be obtained then this is also worth using.

The earthplanes on both sides of the board are joined at many points b wire "worms" which are short lengths of 20–24swg wire (eg scrap com ponent leads) soldered in all the remaining holes that are not cleared on t upper side of the board. These are located all round the edge of the boar as well as in a number of other places. It would be possible to replace t worms round the edges of the board by some thin copper foil, folded ar

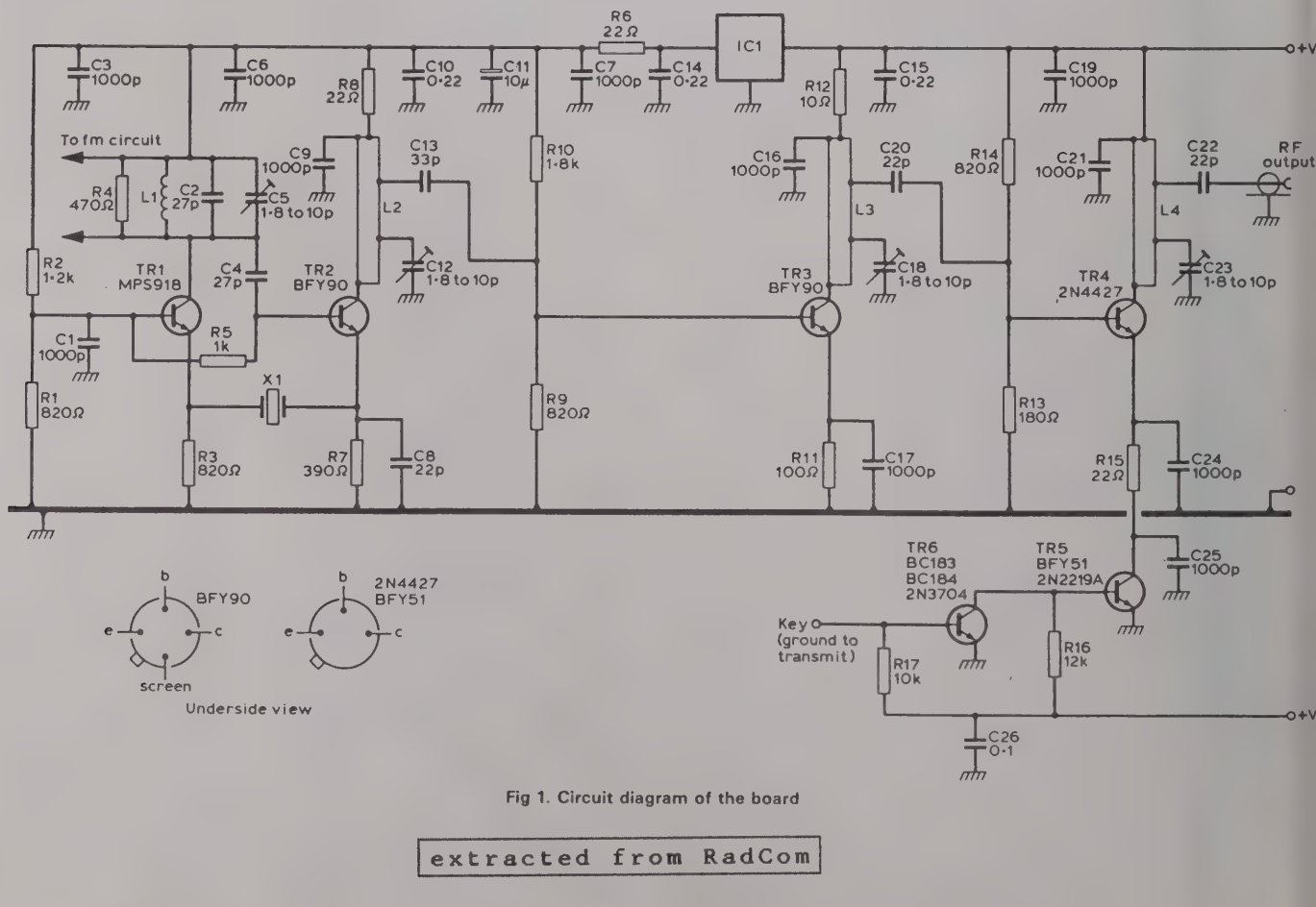


Fig 1. Circuit diagram of the board

extracted from RadCom

soldered along the edges of the board, connecting the two earthplanes. Do not forget the two wire links on the board connecting L1 and L4 to the positive supply. Holes are provided for either the 2- or 3-pin types of trimmer capacitor, which should be mounted so that their movable vanes are connected to the earthy side of the tuned circuits to minimize stray capacitance coupling and loading effects when trimming tools are used.

The output can be taken either by small-diameter coaxial cable directly from the board, with the cable outer soldered carefully to the top earthplane (Fig 6), or from an SMB, SMC (Conhex) or SMA socket on the board.

Provision has been made on the board for a TO220 voltage regulator for the oscillator supply, which reduces the chirp on the note considerably if the board is to be A1 keyed. These regulators generally require an input voltage at least 2·5V greater than the regulated output voltage, so if a 12V one is used (eg 7812, LM34OT12), the supply to the board should be at least 14·5V. If only a 12V supply is available, as would usually be the case when operating portable, either use an 8V regulator (7808)—though this may reduce the output slightly—or omit the regulator altogether and insert a wire link between the two outermost regulator holes.

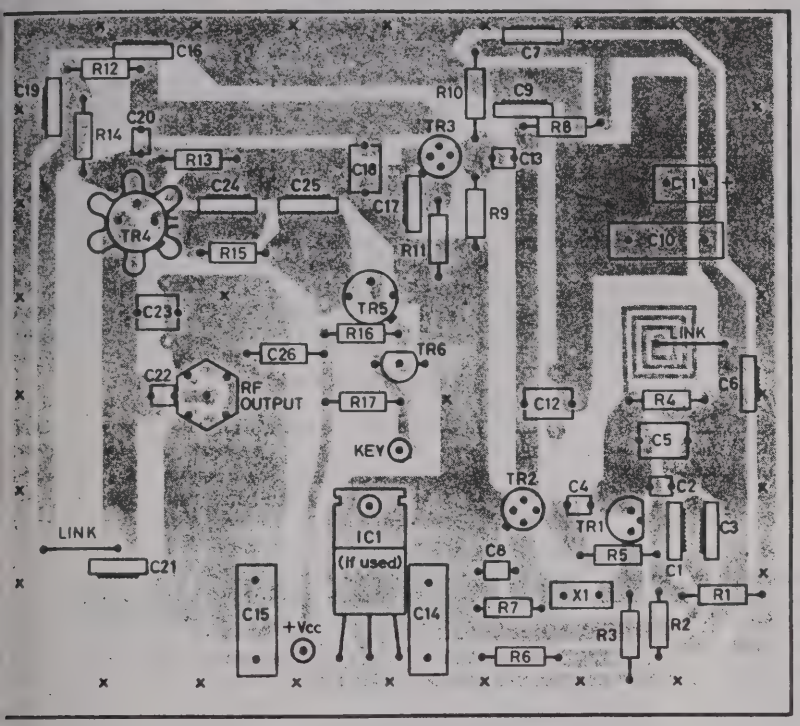


Fig 3. Component layout (x indicates worm joining top and bottom earthplanes)

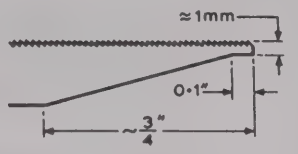


Fig 4. Slot-cutting tool

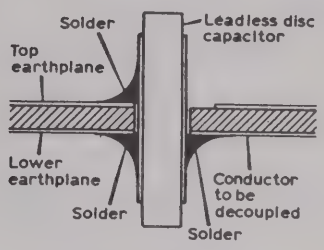


Fig 5. Soldering of leadless disc capacitors

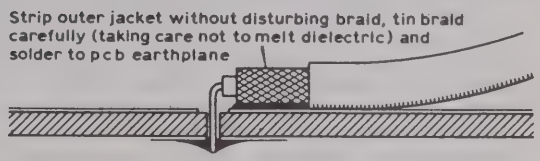
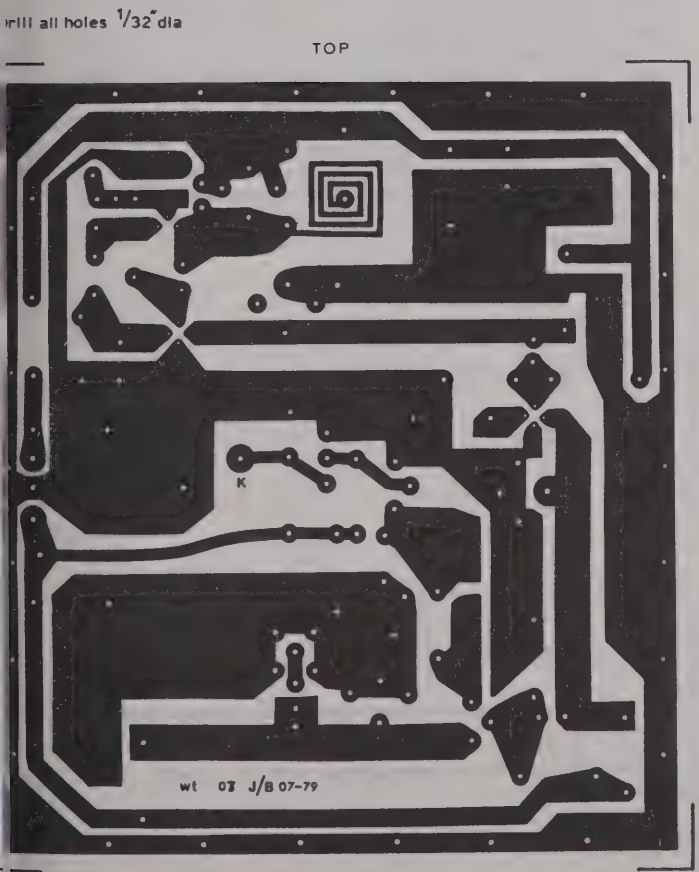

Strip outer jacket without disturbing braid, tin braid carefully (taking care not to melt dielectric) and solder to pcb earthplane

Fig 6. Soldering of output cable to board



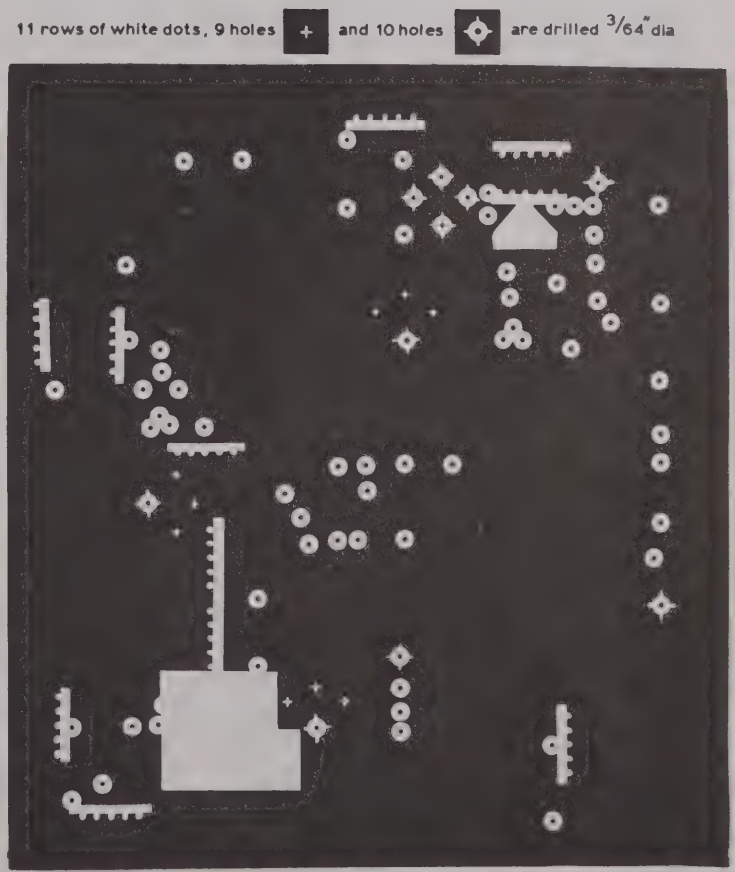


Fig 2. Board artwork

10.1.1 page 3

Components list

MAIN BOARD

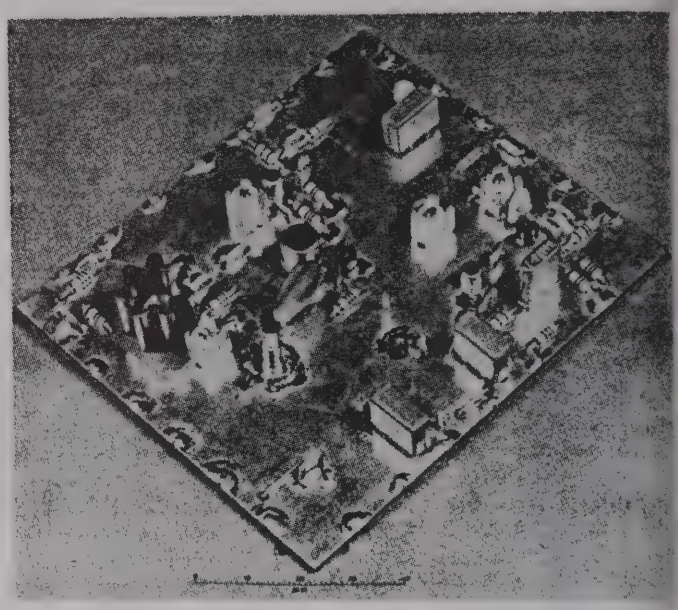
R1, 3, 9, 14	820Ω	R11	100Ω
R2	1·2kΩ	R12	10Ω
R4	470Ω	R13	180Ω
R5	1kΩ	R15	22Ω
R6, 8	22Ω	R16	12kΩ
R7	390Ω	R17	10kΩ
R10	1·8kΩ		

All resistors 0·25W TR4 2 per cent (Electrosil, metal oxide)

C1, 3, 6, 7, 9, 16, 17, 19, 21, 24, 25	1,000pF Leadless disc (Steatite 1,000pF/80/20 TEFK7 400V recommended)
C2, 4	27pF subminiature ceramic disc
C5, 12, 18, 23	1·8-10pF film trimmer (Mullard 809-05002 or Radiospares 125-648)
C8, 20 22	22pF subminiature ceramic disc
C10, 14, 15	0·22μF polyester
C11	10μF tantalum
C13	33pF subminiature ceramic disc
C26	0·1μF ceramic disc

TR1	MPS918	TR5	BFY51 or 2N2219A
TR2, 3	BFY90	TR6	BC183 or BC184 or 2N3704
TR4	2N4427		

IC1 (if used)	LM340T12 (National) or 7812 (Radiospares) 12V or 7808 (Radiospares) 8V
X1	HC18/U or HC25/U fifth overtone crystal
L1 to L4	Printed microstrip lines
Heatsink	TO5 (TR4)
PCB mounting socket	SMA, SMB or SMC



The assembled board

Decoupling, screening and supply regulation

The board is designed to fit in an STC (ITT) 116 by 90mm diecast box (part number 46R CS00 043 A00) with the internal ribs filed away to fit. The board should be mounted, by bolts in its corners, about 5mm off the bottom of the box to avoid excessive damping of the lines by the box, and to ensure that the heatsink on TR4 does not touch the lid of the box. This, or similar screening, is recommended for best results, as any feedback from subsequent stages to the oscillator can degrade the note. Whatever type of power supply is used, it is essential to prevent any rf or af noise from reaching the board on the supply, as this can cause either a noisy carrier, or sidebands on the output, so all connections, eg supply, keying etc, must be thoroughly decoupled. A feedthrough where the leads pass through the box, with a small 10–100μH rfc outside the box, in series with the supply, should be adequate, although a Filtercon is preferable. If possible check with a wide-bandwidth oscilloscope that any voltage regulator circuits are not oscillating—possibly at quite a low level. This can occur at anywhere from audio to tens of megahertz.

The board should be earthed to the box at several places, and the lid should be screwed firmly in place, as any intermittent contact here may cause jumps in frequency. The outer conductor of the rf output should be earthed where it passes through the box, eg by mounting a socket on the box. In general the amount of effort that must be put into these precautions depends upon the factor by which the frequency is to be multiplied.

The on-board regulator, IC1, is probably only worth using if it is intended to use the A1 keying facility, or if a +15V supply is available. Without the regulator the board will operate satisfactorily from a supply as low as 11V, with little drop in output.

Modulation facilities

The pa stage can be on/off keyed by TR5 and TR6. However, this pul the oscillator frequency slightly, and the keying chirp at 10GHz is ur acceptable unless the oscillator voltage regulator (IC1) is used, and eve then it is still noticeable. Frequency shift keying is the preferred method c keying. If the A1 keying facility is not required, TR5, TR6 and the associated components should be omitted, and the end of R15 previousl connected to TR5 grounded directly.

Frequency modulation or fsk can be produced using the circuit in Fig in which a varicap diode varies the capacitance across L1. The value of C must then be reduced to maintain the total capacitance at 27pF. Th BB105B (C=7pF at −5V) with C2=18pF should give a deviation o about 2kHz/V at 10GHz. Wider deviation could be obtained by using a higher capacitance diode such as the BB110G, or a hyper-abrupt type whicl has a larger capacitance swing, eg the Alpha DKV6520 [6].

Inputs are provided for audio at low impedance (source impedance les than 1kΩ), or cw where earthing the input shifts the carrier about 1kHz h at 10GHz; the shift can be altered by changing the 47kΩ resistor.

The components to the right-hand side of the dotted line should b mounted directly across the tuned circuit. The rest can be mounted in plac of the A1 keying circuit if this is not required, but in any case they shoul be mounted on the board to avoid earth loop problems. Efficient filterin is necessary on both inputs to prevent any stray af or rf noise causing un wanted sidebands on the output.

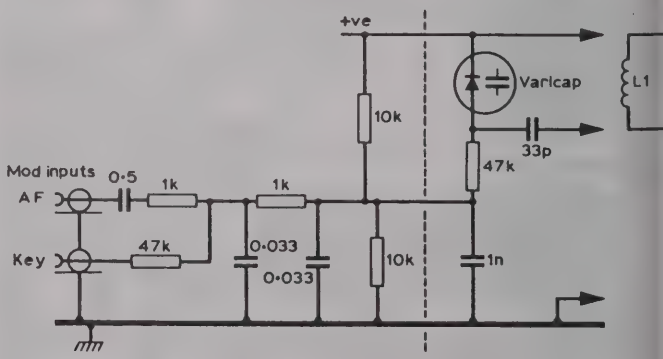


Fig 7. Network for fm/fsk modulation

Alignment

The tuning-up procedure for this board requires the use of some form of output indicator. An hf/vhf swr meter connected to the board via a short length of coaxial cable can be used, set initially on its most sensitive range, and with a 50Ω load or resistor on its output socket. It is recommended, however, that the power meter described below is used, as it is a simple, yet remarkably accurate way of measuring power in the range 100mW to 2W, up to at least 500MHz. Constructional details are given in Fig 8.

The meter indication, V, is the peak rf voltage across the load minus the forward voltage drop across the diode. This is divided by $\sqrt{2}$ to give the rms value, and the power is calculated from this using the equation

$$P = \frac{(V_{rms})^2}{R}$$

For germanium diodes, such as the OA47, the relationship will therefore be

$$P = \frac{(V + 0{\cdot}25)^2}{100};$$

while for silicon diodes, such as IN914, IN916, IN4149, it is

$$P = \frac{(V + 0{\cdot}7)^2}{100}.$$

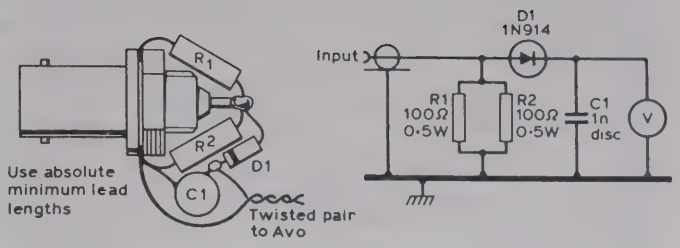


Fig 8. Construction of power meter on bnc socket

The results from this meter using a IN914 at 500MHz are compared with the actual power in Fig 9, and show the high accuracy obtainable.

Apply power to the board, and check the current drawn (approximately 150mA). For a 96MHz crystal, preset the trimmers to the following positions: C5 about 50 per cent meshed, C12 about 40 per cent meshed, C18 about 80 per cent meshed and C23 about 50 per cent meshed. Some output should be detected, and this is maximized by peaking the trimmers.

It should not be possible to tune up on the wrong frequency using the component values specified, but if in doubt check with an absorption wavemeter. The position of C5 should not be too critical as regards output power, and may be used to trim the frequency. If crystal frequencies outside the range 90–100MHz are used it may be necessary to alter the value of C2 so that as C5 is adjusted over about half its travel, the power output stays constant but the frequency pulls smoothly. If necessary the frequency can be raised slightly by putting a 5–20pF trimmer in series with the crystal, but this may reduce the power output or stability, and in general it is preferable to put up with a small frequency offset (which can be calibrated out) rather than compromise on stability.

When the board is tuned up it should produce a comfortable 100mW rf output. It may be possible to increase this slightly by optimizing certain components (eg R10, C8, C22) but if more power is needed it is preferable to add an extra stage, rather than risk detracting from the stability, low noise sidebands and reliability of the oscillator board.

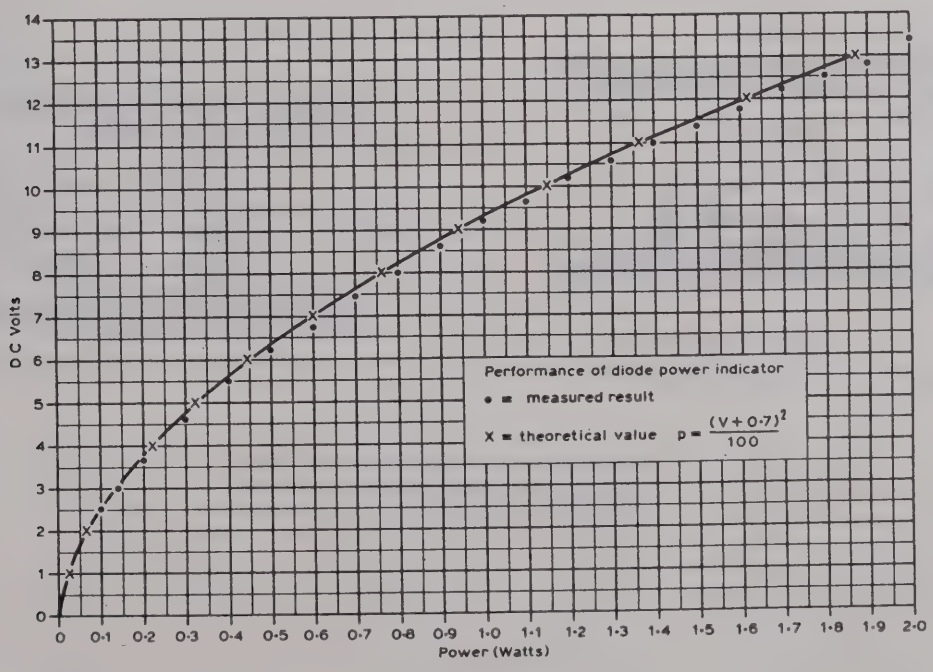


Fig 9. Power meter performance

10.1.3 Stabilised Low Voltage Drop Regulator

For best frequency stability the whole board must run from a stabilized supply, and it is desirable that this should be the same voltage whether for portable or fixed operation, so that the frequency calibration is maintained. This is particularly important when using car batteries, as their voltage varies somewhat according to the state of charge, current drawn, etc (perhaps as much as 11·5V to 16V), and the same is true of some other types of battery. The circuit shown in Fig 10 makes the oscillator well and truly insensitive to supply voltage variations, and provides overvoltage and reverse supply protection. It will regulate down to a minimum voltage drop of approximately 300mV (governed by the Vce_{sat} of the transistor). A pcb design is given in Fig 11, and the board layout in Fig 12.

The relay used for reverse supply protection (Radiospares 348-510) is rated at 1A 28V, so this circuit can be used (and is highly recommended) for all pieces of gear that need a regulated supply up to this rating, and are likely to be used portable. RV1 sets the regulated output voltage; something like 11·2V is suggested as a reasonable value that will guarantee a regulated output even at the lowest supply voltages. The BD132 transistor is bolted to a suitable heatsink, with the usual insulating washer.

LOW-VOLTAGE DROP REGULATOR BOARD

R1	4·7kΩ	C1, 3	10μF 25V
R2	15kΩ	C2, 4	1nF disc
R3	1kΩ	C5	100μF 25V
R4	3·3kΩ	IC1	741
R5, 6	18kΩ	TR1	BD132
D1, 2, 4	IN914, IN916 or similar Si diode	RLA1	Radiospares 348-510
D3, 5	4·7V zener		
D6	5·6V zener		
RV1	5kΩ Cermet D (Radiospares 184 502)		

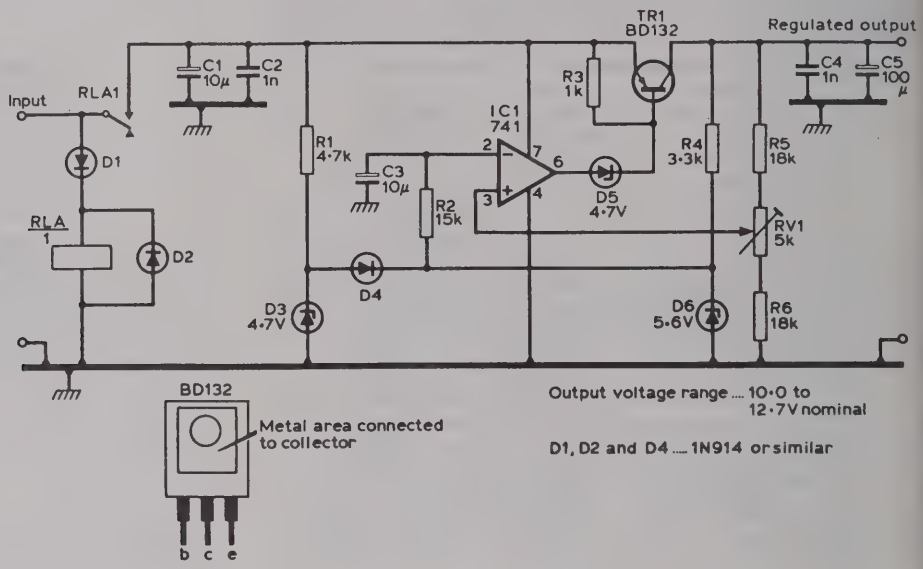


Fig 10. Low voltage-drop regulator

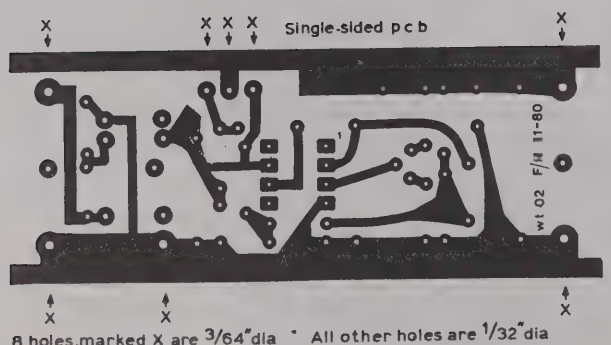


Fig 11. Board artwork for low voltage-drop regulator

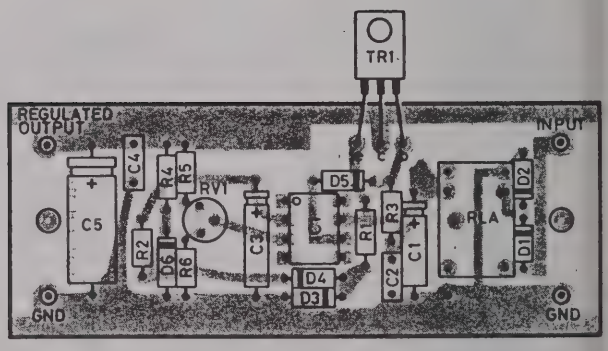


Fig 12. Component layout for low voltage-drop regulator

extracted from RadCom

10.1.4 Performance

The following results give typical performance figures measured on the board shown in the photograph:

Supply voltage (V)	Output power (mW)
11	225
12	270
13·5	325
15 (without IC1)	350
15 (with IC1)	330

The spectrum of the output is shown in Fig 13. This varies very little with supply voltage. Unwanted products are better than 35dB down on the main carrier.

Frequency stability (at 10GHz)

Variation of frequency with supply voltage:
- without IC1 — 5kHz/V
- with IC1 — <100Hz/V (once regulating)
- with regulator of Fig 10 — hardly detectable

Variation of frequency with temperature: 2kHz/°C (this depends somewhat on the crystal)

Applications

Most of the boards that have been built so far have been used on 10GHz to provide a local oscillator for the G3JVL 10GHz transverter [2] in conjunction with the G8DEK step-recovery multiplier design [3]. Most of the comments in this article refer to measures required to give a good note at this frequency. For this application a 94·666MHz crystal is used to give an output at 378·666MHz, for a final local oscillator frequency of 10·224GHz. An output power of about 2·5W is needed, so some form of additional amplifier is required. This should not be mounted in the same box as the board.

Amplifiers that have been used with success for this include the Mullard BGY22 module (which requires an attenuator [4] at its input to reduce the power from the board to approximately 50mW) and the Wood & Douglas MD3PA. While the former is very simple and convenient, the latter has rather better spectral purity. The Mullard BGY22/BGY23 combination or the Wood & Douglas MD10PA [5] are also recommended as higher power amplifiers for transmitter-type applications.

Crystal frequencies and rf output frequencies for various applications are given in Table 1.

Conclusion

Although the board is quite simple to construct and align, there is no compromise in performance. Its versatility should eliminate all problems of building a new local oscillator strip design for each new microwave project.

References

[1] Steatite-Roederstein Ltd, Hagley House, Hagley Road, Birmingham B16 8QW. (**Note:** £50 minimum order charge.

[2] "G3JVL 10GHz transverter". *Rad Com* January and April 1979, pp41, 342, April 1980, pp372–5, February 1981, p146.

[3] "G8DEK BXY41E 10GHz step-recovery diode multiplier". *Rad Com* March 1976, p202.

[4] "Design and construction of simple attenuators". *Rad Com* March 1979, p239.

[5] "Equipment review—Wood & Douglas 384MHz MDO5T microwave drive source and MD10PA power amplifier kits". *Rad Com* June/July 1980, pp650–2.

[6] Alpha Industries, RMC House, Station Road, Witney, Oxon OX8 6BP.

□

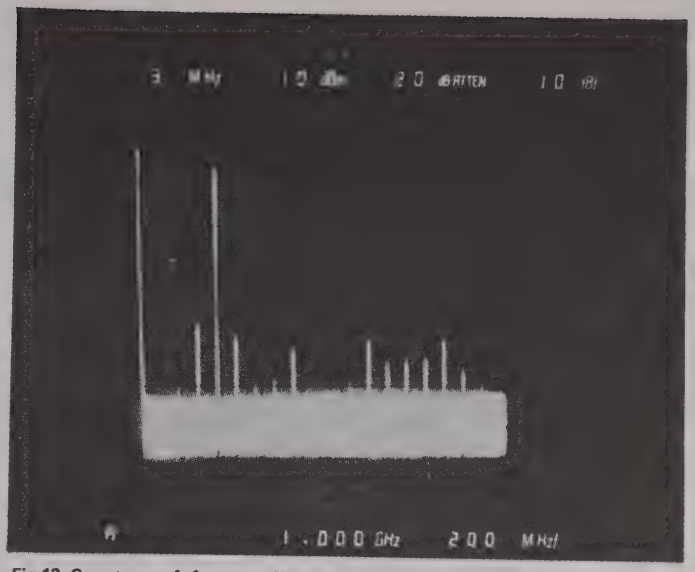


Fig 13. Spectrum of rf output. Horizontal 0–1·8GHz, 200MHz/div. Vertical: top of screen = +30dBm, 10dB/div. Analyser bandwidth = 3MHz

Table 1. RF output frequencies and crystal frequencies for various applications of the board

Application	RF output frequency (MHz)	Crystal frequency (MHz)
Microwave cw/fsk/fm tx	384	96
1·3GHz converter/transverter (144MHz i.f.)	384	96
2·3GHz converter/transverter (144MHz i.f.)	360	90
3·4GHz converter/transverter (144MHz i.f.)	368	92
5·7GHz converter/transverter (144MHz i.f.)	374·4	93·6
10GHz converter/transverter (144MHz i.f.)	378·666	94·666
24GHz converter/transverter (144MHz i.f.)	381·714	95·4286

Temperature Compensation for the 384MHz Oscillator Board - G3YGF/G3JVL

This circuit reduces the effect of temperature on the oscillator frequency by a factor of at least ten by compensating for thermal drift.

The temperature coefficient has been measured on a number of these boards, and seems to be about +- 0.2ppm per degree C. This results in a frequency shift of about 40kHz on 10GHz for the extremes found at a home station, eg 10 - 30 deg C. Typical temperature coefficients for AT-cut crystals range from 0 to +- 1ppm per degree. Thus the main cause of the drift when using standard spec crystals seems to be the crystal itself, and the rest of the circuit has much less effect. Because the crystal is very loosely coupled to the circuit, - as is desirable in a highly stable oscillator - it is very difficult to perform compensation using normal temperature sensitive capacitors as very high rates of change are required.

This circuit uses a thermistor glued to the crystal can in a potential divider circuit to produce a voltage that varies with crystal temperature. This voltage is then used to drive a varicap diode connected across part of the oscillator circuit. This produces a capacitor with a temperature coefficient effectively variable up to +- 20,000ppm. The circuit is designed to be fairly immune to supply voltage variations, but should still be run off the same stabilised supply as the oscillator.

The circuit can be built in 3-D on a small piece of double sided PCB and will fit on some of the spare space on top of the oscillator board. The thermistor leads will reach to the crystal, and the +11V and lead to the varactor can go through holes drilled in the PCB.

If the FSK or FM mod inputs are required as well, then the output of the filters shown in the article can be fed into point A on the varactor circuit, and the 1uF capacitor removed.

The voltage from the thermistor varies linearly with temperature over the range 0 - 60 degrees, and is then passed through an op amp with variable gain and offset so that the mean voltage on the varactor and thus the centre frequency, and also the rate of change of voltage with temperature can be set independantly. As shown, the circuit compensates for oscillators with a negative temperature coefficient; to deal with those with a positive coefficient, just swop over the thermistor and R3.

The characteristic of the DKV6533A is fairly linear over the range 5 - 10V, so the output should be set to give 7.5V across the varactor at ambient temperature, and the gain to about unity, initially (RV2=100k). They are available from Alpha Industries, RMC House, Station Rd, Witney, Oxon, OX8 6BP for 2.46 each. The BB110G has the correct change in capacitance, but is not as linear over the 5-10V range, and is a poor substitute as it will not give as accurate a correction over the full temperature range.

For best results the gain preset will have to be set up individually on each board. This requires a means of measuring small changes in frequency accurately, although as the tests can be done fairly quickly, and the reference left in a constant environment while the board under test is subjected to extremes of temperature, the demands on the stability of the reference are not that high.

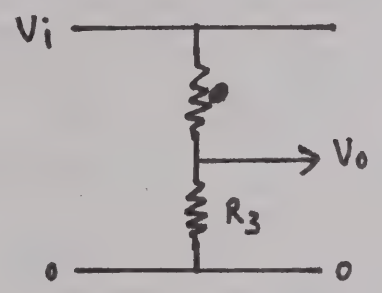
With the gain set near unity, measure the frequency shift over a range of temperatures, eg 5-10 deg=bottom of a fridge, 20 deg= ambient, 40deg=near a heater. Allow plenty of time (20-30 mins) for the temperature inside the oscillator box to reach equilibrium. The change in gain needed can then be either calculated, or guessed, and the process repeated. It may be found that the compensation required is non linear, if the error is different at high and low temperatures. If this is the case, the value of R3 can be changed to decrease the compensation at the appropriate end of the temperature range, (see table) and the process repeated. Once the approximate value needed for RV2 is known, it it probably worth altering

the value of R2 so that RV2 is near max resistance, or replacing it with a fixed resistor and a smaller value preset in series to enable finer adjustments to be made. As any movement of the presets will affect the frequency, it might worth using better quality ceramic types, or ten turn ones.

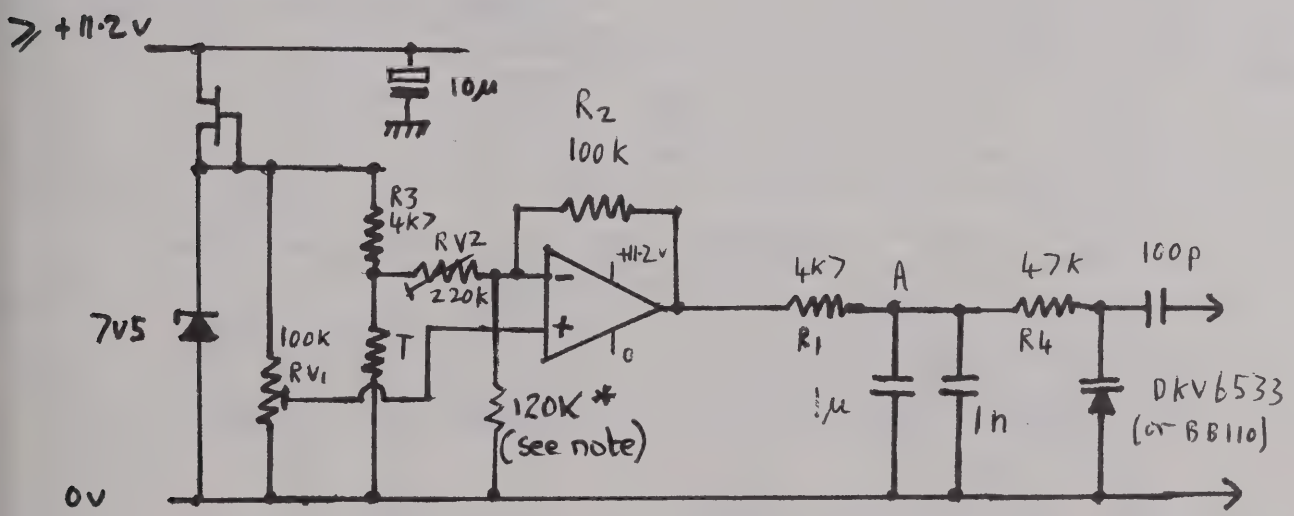
Needless to say, no other factors should be able to affect the frequency during this test; the board should be mounted rigidly in a well screened box, with no holes where draughts could enter, and run from a stabilised power supply.

By this means it should be possible to achieve a temperature coefficient of 0.01ppm per degree, a drift of 1kHz at 10GHz over the normal operating range. Having got the oscillator this stable, the only really reliable way of measuring the frequency accurately is to use an off air standard, such as Rugby or Droitwich. Be sceptical about the accuracy of commercial counters! If you wish to achieve still higher accuracy, then this compensated oscillator can be locked to one of the off air standards, using a loop with a very long time constant to prevent the modulation pulling the frequency. It is vital to have an extremely stable oscillator before trying to phase lock it!

Temp Degs C	Vo/Vi R3= 2k	5k	10k	Res of thermistor
0	.13	.23	.43	13k
10	.18	.31	.53	9k
20	.26	.41	.64	5k7
30	.34	.51	.72	3k9
40	.42	.60	.79	2k7
50	.54	.70	.85	1k7
60	.67	.80	.91	1k0



Characteristics of varicaps:
DKV6533 5V 8pf 10V 4pf
BB110G 5V 22pf 10V 17pf



T 5k thermistor at 25°C LF355 (any Fet op amp)

RV1 Set centre frequency

RV2 Set gain (20K-2M2)

R3 set Law (>5k reduces effect at high temp, <5k reduces effect at low temp)

* Note: The resistance of RV2 can lie between about 20k and several Mohm; if a greater change of gain is needed, R2 can be changed, but then the 120k resistor must be changed in the same ratio.

Upper frequency limit of 'Microwave Committee' uhf sources - G4CNV

Several people have used this basic design at frequencies quite a long way from the original 384MHz (4 x 96MHz). Certainly, with a 108MHz crystal it makes quite a useful 432MHz source; also G3JVL has for a long time used a higher basic crystal frequency to reduce the multiplication factor from 4 to 3 (eg for 384MHz a 128MHz crystal would be used), and obtained higher efficiency as a result.

Faced with a requirement for low-noise sources at several frequencies between 500 and 600MHz, an attempt was made to see if the basic design could be extended.

The major factor influencing this is the maximum resonant frequency of the tuned circuits. As it stands, the maximum frequency of the basic design seems to be about 500MHz (the original Plessey AMETS design used 101MHz crystals to give an output around 404MHz).

Referring to the circuit diagram opposite, reproduced from the Radcom article (October '81), it will be seen that there are four tuned circuits:

L1//C2,C5; L2//C12; L3//C18; and L4//C23.

The first, L1//C2,C5 resonates at the basic crystal (usually 7th overtone) frequency. It is fairly easy to raise the resonant frequency of this to beyond 200MHz by:

(i) reducing or removing C2.

(ii) replacing C5 (Mullard type 809-05002, 1.8 - 10pF, or RS 125-648, 2 - 10pF) by one with a lower minimum capacitance and lower swing. Mullard type 809-05001, 1 - 3.5pF is suitable. A lower capacitance swing makes adjustment easier - removing C2 makes the variable part of the total capacitance much larger, and it will be found that crystal locking of the oscillator only occurs over a portion of the adjustment range. It is therefore preferable that the adjustment range should not be too great.

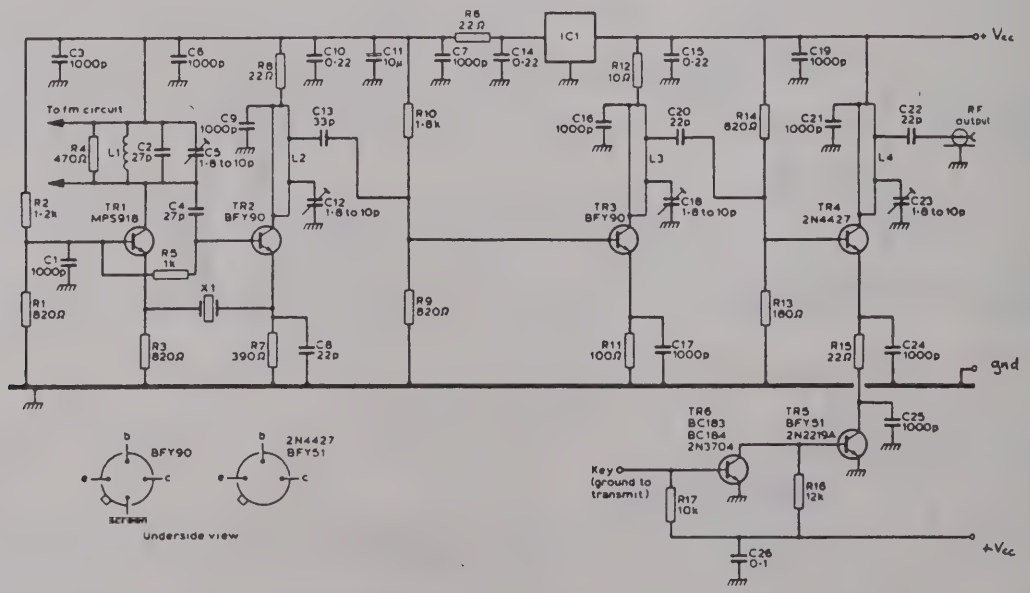
(iii) shorting out a portion of L1 at the centre with solder.

Higher-frequency crystals are more expensive, but it is generally possible to get a particular crystal to oscillate on a higher overtone. Note that only oscillation on odd overtones is possible, and that the frequency ratio will not be exact (though very close). Thus a 7th overtone 96MHz crystal would also be able to oscillate at its 9th overtone, (96/7) * 9 = 123.428MHz.

The maximum resonant frequency of the remaining three tuned circuits can be raised by substituting lower-minimum-value trimmers as in (ii) above. When making adjustments, bear in mind that simply being able to peak the output level with a trimmer capacitor does not necessarily indicate correct resonance - it should be possible to find <u>two</u> such positions either side of minimum capacitance. The fourth tuned circuit, L4//C23, seems to have the lowest maximum resonant frequency, which is about 580MHz. Since the performance of the transistors (especially the 2N4427) falls off quite rapidly with frequency, a more suitable output device would be the BFR96, though as this comes in a different package, it would have to be soldered in position underneath the board. The lower collector capacitance of this device would raise the maximum resonant frequency of L4//C23.

G3JVL has in fact been able to use the original AMETS board to give output at 900MHz, by physically shortening the lines. On the original design, which is a positive-ground circuit, this is quite easy, since the upper ends of the lines are actually grounded rather than just decoupled. Shortening the lines on the negative-earth version would still be feasible, but would require rather drastic surgery to relocate the leadless disc capacitors (C9, C16, and C21). The table opposite gives some idea of the results obtained at various frequencies using a standard board with the usual transistors, and substituting lower-value trimmer capacitors where appropriate.

Mullard BGY22c amplifier modules are specified to work from 380 to 512MHz giving 17dB gain. Several were tried at 532MHz, and all gave about 14dB gain, so would still be usable. Note, though, that they still have their full gain at the upper subharmonic frequencies, so will degrade the relative level of these subharmonics.



o/p freq (MHz)	384	404	464	524	594
xtal freq (MHz)	96	101	116	174.666	198
mult factor	x4	x4	x4	x3	x3
P_{out}	+24dBm (250mW)	+20dBm (100mW)	+22dBm (160mW)	+21dBm (130mW)	+17dBm (50mW)
L_1	as is	as is	as is	1/2t shorted	1t shorted
C_2	22p	22p	8p2	zero	zero
C_5	1.8-10p, 70%	1.8-10p, 30%	1.8-10p, 90%	1-3.5p, 60%	1-3.5p, 50%
C_{12}	1.8-10p, 50%	1.8-10p, 50%	1.8-10p, 20%	1-3.5p, 60%	1-3.5p, 10%
C_{18}	1.8-10p, 70%	1.8-10p, 70%	1.8-10p, 50%	1.8-10p, 10%	1-3.5p, 50%
C_{23}	1.8-10p, 70%	1.8-10p, 40%	1.8-10p, 20%	1-3.5p, 40%	1-3.5p, minimum

10.1.7 Additional Notes

G4FRE has found that the Birkett BGY22C although specified as 420- 480MHz can still give 2W at 378MHz. He has also done some measurements on the 384MHz oscillator board, and says that the PA keying circuit only reduces the output by 13dB. (I would not recommend using this for A1 keying though, as the frequency does pull slightly, and F1 would be a much nicer method of keying it -G3YGF).

Microwave Committee 384MHz oscillator boards
GM3FYB suggests replacing R9 by a 1k pot, to give control of output from 0 to full value (350mW in his case). In some earlier versions, R22 was given as 15ohms. This can lead to the PA getting excessively hot whilst only slightly increasing the power output. A value of 22 or 27 ohms is reccommended.

10.1.8 Oscillator Quality on 10 GHz:

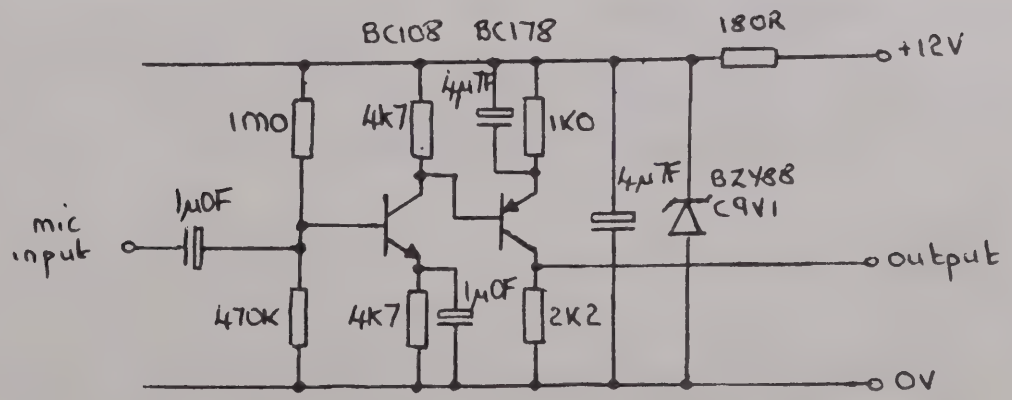
Some vital points for all microwave drive sources are:

1. Decouple the supply very thoroughly - any rf or af on the supply can produce very bad FM noise sidebands at 10 GHz.
2. Make all components in the oscillator mechanically rigid, including the fixing of the board to the box, and the lid on the box. Bad contacts here can cause the frequency to jump.

10.1.9 Mic Preamp

The following is a circuit used by GW4LXO for feeding a 384 MHz source for NBFM. The output is connected to a filter mounted on the 384 board. No doubt the circuit could be put to use as part of a Gunn diode psu also.

MIC PREAMP



VHF Comms Local Oscilators

G4FRE has sent details of local oscillators built by himself and other people in the Ipswich area from designs in VHF Comms:

DC8NR006 in 2/76 p67 — Output 40mW at 1152MHz

This design was not found to be very stable. The adjustment of C10 is particularly critical. A BB105B seemed to work better than the BA149. Some adjustment of the dimensions of L7 were found to be necessary in both versions built. The output powers at best were about 20mW.

DC0DA003 in 4/76 p202 — Output 5mW at 2160MHz

This design worked well up to T6 where 50mW of 540MHz was measured. A problem area was then encountered. Two versions have been built, one with C11 a glass JFD trimmer, the other with C11 a brass disc on an M3 screw. Neither versions worked efficiently even after much experimentation with the line diameter, trough sizes and multipier diodes. The best output measured, with a BXY27 diode, 7mm tube and 9mm tuning disc, was 4mW.

DC0DA005 in 1/78 p18 — Output 200mW at 1152MHz

The main problem encountered with this design was the general non availability of C1-12. Various surplus surplus transistors were substituted - 61387, 218BLY etc. With these transistors the output has ranged from 50mW to 100mW. Suprisingly it was found that the BB105B worked well as a tripler to 1152MHz in this (and other) design. A BB105B has even produced 350mW output at 1296MHz when tripling. The mask on p21 is shown full size.

DC8UG001 in 2/78 p66 — Output 800mW at 1152MHz

This consists of a 384MHz source of approx 1.5W (DC8UG001) and a transistor tripler (DC8UG002).

1. Source. Again the availability of C1-12 was found a problem. As before, other surplus devices were used producing in excess of 1W. The pcb mask on p72 is shown full size, but beware of misinterpreting symbols on this layout. A closed symbol in VHFC normally represents an earthed connection. However T1 emmitter is not earthed (as correctly shown on the circuit diagram on p67). Conversely the 5p6 capacitor shown connected on the layout between T1 emmitter and collector is really connected between T1 collector and earth (as shown on p67).
2. Tripler. Surplus transistors were used to replace C3-12. The design worked with a spectrally clean output due to the 2 section filter. If only the 0.8W output port is to be used the unused port must be terminated with 50 ohm (see p127 of the same issue).

DF8QK002 in 2/78 p125 — Output 10+5mW at 1268MHz

The design was published twice. The latter version has altered inductor dimensions and works better. The oscillator uses 70.444MHz crystals but 96 and 105.666MHz crystals have been successfully used in an oscillator circuit similar to the DC0DA design. Using a BFY90 for T1 to T4 and a BFR34a for T6 then 20mW and 7mW outputs were produced. The pcb mask is not shown full size (4/78 p242).

DC0DA006 in 3/78 p66 — Output 60mW at 1104MHz

The pcb mask is shown full size. Using a BFY90 for T2 to T4, a BFR91 for T5 and a BFW16a for T6 the output was 50mW. Using a 2N3866 for T6 was not as good, giving only 20mW.

Cystal Oscillators: G8ADP has built 5 boards based on the VHF Comms DC0DA 005, and finds that they all come out 10 to 20 kHz higher than expected at 384 MHz. Several MM oscillators use a 2 pF capacitor acsoss the emitter-collector of their oscillators so the frequency can be pulled onto the correct value. Stability does not seem to be good enough for microwave purposes.

Section 11 Filters

Design Calculations for Waveguide Bandpass Filters - G4KNZ

Many people will be familiar with the filters used in the "JVL" image recovery mixer on 10GHz. They are 3 pole bandpass filters realised by dividing up a section of waveguide into three cavities, using iris plates. The holes in these plates provide coupling between the cavities.

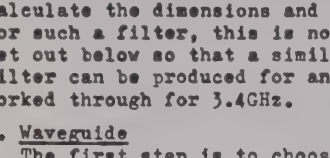
More recently, G3JVL and others have produced designs for 1.3GHz (repeater duplexing filters) and 5.7GHz. While at first sight it might seem a horrifying task to calculate the dimensions and iris sizes from scratch for such a filter, this is not so. The steps needed are set out below so that a similar bandpass waveguide filter can be produced for any band, and an example is worked through for 3.4GHz.

1. Waveguide

The first step is to choose a suitable waveguide (see tables in the VHF/UHF manual, for example) and determine the inside dimensions. These are referred to as a and b in the text (a being the wider dimension). Usually the choice is dictated by what is available - there will be one standard guide available for the band of interest.
eg for 3.4GHz WG11: a=60.25mm
b=28.50mm

2. Centre Frequency and Bandwidth

Establish the frequency and bandwidth required. The lower and upper frequencies the filter is required to pass can then be determined. These will be referred to as f_1 and f_2 . The centre frequency will be referred to as f_o. Note that the bandwidth is the 3dB bandwidth, so that at f_1 and f_2 the insertion loss will be 3dB more than at f_o.



eg f_o=3456MHz
& B/W=20MHz

Then f_1=3446MHz
and f_2=3466MHz

In practice the insertion loss of the filter at f_o will not be zero. In general, for a given number of cavities in a particular waveguide, decreasing the bandwidth will increase the insertion loss. It is possible to calculate the theoretical insertion loss, but a rough guide is that for a ratio of Bandwidth/Centre Frequency of 0.002 a loss of the order of 1dB can be expected. In the above example, $B/W \div f_o = 20 \div 3456$ = approx 0.006. This is about 3 times the figure of 0.002 and so a loss well below 1dB should be expected. The situation is not as simple as this and we hope to have more on calculating these losses in the future. For the time being, assume that a minimum of 0.002 should be used for the ratio of bandwidth to frequency (called fractional bandwidth). This ratio corresponds to about 20MHz bandwidth at 10 GHz, and 2.5MHz bandwidth at 1.3GHz. It is worth noting that another problem with unduly narrow filters is that they will be more critical to align.

3. Waveguide Wavelengths

The wavelengths of frequencies f_1 and f_2 in the chosen guide need to be calculated from the formula:

$$\lambda_g = \frac{299.7}{\sqrt{F^2 - 22472/a^2}}$$

where F is measured in GHz, and a in mm.
Thus for the example given:
At frequency f_1 = 3.446GHz, λ_{g1} = 125.7mm
f_2 = 3.466GHz, λ_{g2} = 124.2mm

Then calculate the centre wavelength:

$$\lambda_{go} = \frac{\lambda_{g1} + \lambda_{g2}}{2}$$

Thus for the example given:

$$\lambda_{go} = \frac{125.7 + 124.2}{2} = 124.95\text{mm}$$

4. Fractional Bandwidth

Calculate the fractional bandwidth in terms of the wavelength:

Fractional Bandwidth, $W_\lambda = \frac{\lambda_{g1} - \lambda_{g2}}{\lambda_{go}}$

Thus for the example given:

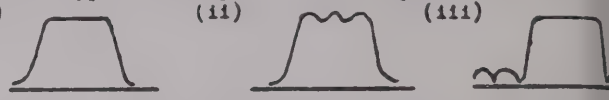
$$W_\lambda = \frac{125.7 - 124.2}{124.95} = 0.0120$$

Note that this is not the same as the fractional bandwidth in terms of frequency, $(f_2 - f_1)/f_o$. For the example given, this would work out at 0.0056, which is approximately half W_λ. This is quite in order Be sure, therefore, to follow the calculations exactly

5. Type of Filter Response required

For anyone not very familiar with filters, there ar various types of bandpass filter responses such as:

(i) (ii) (iii)



It is possible to design the filter to have the particular response required. The design here will be for a response of type (ii). See one of the many books on filter design for more information on other types. This type of filter chosen can be designed for a given amount of ripple in the passband eg for 0.1dB ripple o 1.0dB ripple. Basically, accepting more ripple in the passband gives a better rejection outside the passband A reasonable compromise is 0.2dB ripple.

Again see text books for more information on other values of ripple, or more data can be written up here in the future if there is a demand. (This type of filter response is known as the Tchebyscheff response. The filter is to have three poles ie it is to consist of three waveguide cavities. Other numbers of cavities are possible but are not considered here.

6. Normalised Element Values

The desired filter response enables what are known as normalised element values to be chosen from publish tables. In fact the values refer to reactance values for a filter of the required respohse to work at a frequency of 1. These normalised values are simply inserted in to appropriate equations as constants. The are introduced here for the benefit of anyone wishing to use alternative constants for alternative filter responses.

For the type of filter response chosen, the constan required are:

g_o = 1.0000
g_1 = 1.2275
g_2 = 1.1525
g_3 = 1.2275
g_4 = 1.0000

7. Calculate constants for the Filter of Bandwidth W

$$K_{01} = \sqrt{\frac{\pi}{2} \times \frac{W_\lambda}{g_o g_1}}$$

$$K_{12} = \frac{\pi}{2} \times \frac{W_\lambda}{\sqrt{g_1 g_2}}$$

$$K_{23} = \frac{\pi}{2} \times \frac{W_\lambda}{\sqrt{g_2 g_3}}$$

$$K_{34} = \sqrt{\frac{\pi}{2} \times \frac{W_\lambda}{g_3 g_4}}$$

Note that since $g_1 = g_3$ then $K_{12} = K_{23}$ Similarly $K_{01} = K$ and so only the first two constants need to be worked out.
For the example given:

$$K_{01} = \sqrt{\frac{\pi \times 0.0120}{2 \times 1 \times 1.2275}} = 0.1246 = K_{3}$$

$$K_{12} = \frac{\pi \times 0.0120}{2\sqrt{1.2275 \times 1.1525}} = 0.0135 = K_{2}$$

Then evaluate X_{01} to X_{34} which are values of the reactance required at each iris plate position (an iris with hole appears as an inductor), normalised to the waveguide characteristic impedance.

$$X_{01} = \frac{K_{01}}{1 - K_{01}^2}$$

$$X_{12} = \frac{K_{12}}{1 - K_{12}^2}$$

$$X_{23} = \frac{K_{23}}{1 - K_{23}^2}$$

$$X_{34} = \frac{K_{34}}{1 - K_{34}^2}$$

Again in our case there will be no need to work out th last two values, as X_{01} equals X_{34} and X_{12} equals X_{23}. For the example, X_{01} = 0.1266 and X_{12} = 0.0135.

8. Spacing of Irises

We are now able to calculate the length required between the irises to form the cavities. There are thr cavities of electrical length θ_1, θ_2, and θ_3 radians.

$$\theta_1 = \pi - \tfrac{1}{2}(\arctan 2X_{01} + \arctan 2X_{12})$$

$$\theta_2 = \pi - \tfrac{1}{2}(\arctan 2X_{12} + \arctan 2X_{23})$$

$$\theta_3 = \pi - \tfrac{1}{2}(\arctan 2X_{23} + \arctan 2X_{34})$$

Again θ_1 will be found equal to θ_3 (filter symetrical).
For the example given:

$$\theta_1 = \pi - \tfrac{1}{2}(\arctan 2{*}0.1266 + \arctan 2{*}0.0135)$$

ie $\theta_1 = 3.0041$ radians $= \theta_3$

$$\theta_2 = \pi - \tfrac{1}{2}(\arctan 2{*}0.0135 + \arctan 2{*}0.0135)$$

ie $\theta_2 = 3.1146$ radians

The length in radians is now converted to one in mm using the waveguide wavelength λ_{go} which is equal to 2π radians.
So

$$L_1 = \frac{\theta_1}{2\pi} \times \lambda_{go}$$

$$L_2 = \frac{\theta_2}{2\pi} \times \lambda_{go}$$

$$L_3 = \frac{\theta_3}{2\pi} \times \lambda_{go}$$

For the example given

$L_1 = 59.7$mm

$L_2 = 61.9$mm

$L_3 = 59.7$mm

However in a practical filter, it will not be possible to get the lengths exactly right - some tuning will be necessary. So the cavities are made slightly too short so the filter operates on a slightly higher frequency than desired, and then each cavity tuned down to the correct frequency using a tuning screw. It is suggested that the lengths calculated are reduced by 5%.
So for the 3.4GHz filter we use:

$L = 59.7 \times 0.95 = 57$mm (approx)
$L = 61.9 \times 0.95 = 59$ mm
$L = 59.7 \times 0.95 = 57$ mm

9. Tuning Screws

Choose bolt sizes to give a reasonable tuning range without being too critical to set. The following are suggested:

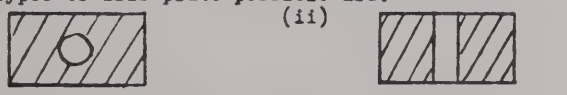
10GHz	use 6BA or M3
5.7GHz	use 4BA or M4
3.4GHz	use 2BA or M5

These should be placed one in the centre of each cavity (see designs for 10GHz JVL image recovery mixer, for example).

10. Holes in Iris Plates

Two types of iris plate possible are:

(i) round hole (ii) rectangular hole



The round hole type has been most commonly used in designs so far and data to calculate round iris hole sizes follows. However, from the calculations so far, it is possible to proceed to use either type as they are only different ways of producing a given shunt inductance across the waveguide. The round holes will be suitable for a filter with a fractional bandwidth similar to that suggested. For a larger bandwidth, the size of the iris holes must be increased. Eventually, the diameter of the holes will be found to be equal to the height of the waveguide, if the bandwidth is increased sufficiently. It will then be necessary to use iris plates with rectangular holes.
Calculate reciprocals of the reactances evaluated in step 7:

$Y_{01} = 1/X_{01}$

$Y_{12} = 1/X_{12}$

$Y_{23} = 1/X_{23}$

$Y_{34} = 1/X_{34}$

For the example, $Y_{01} = 1/0.1266 = 7.9 = Y_{34}$
$Y_{12} = 1/0.0135 = 74.0 = Y_{23}$

Then look up the size of the hole from the graphs below to realise the required reactances. At this stage the aspect ratio of the waveguide is needed. This is the ratio of the broad dimension in the guide to the narrow dimension: $R = a/b$

For the example using WG11 then

$R = 60.25/28.50$

ie $R = 2.11$

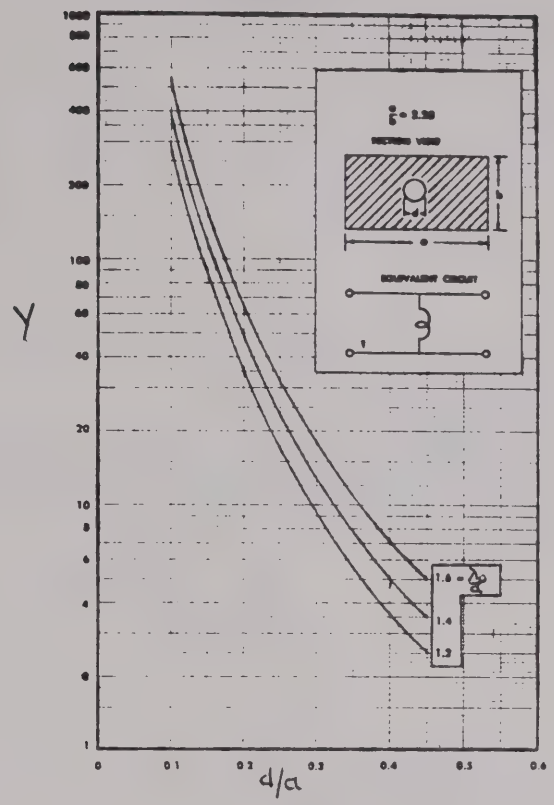
Normally the ratio is between 2 and 2.25, although other values are sometimes used. Graphs are given for ratios of 2.0 and 2.25 to calculate the hole sizes. For an aspect ratio in between these values, use both sets of graphs and interpolate.
Also required at this stage to use the graphs is the ratio λ_o/a, where λ_o is the free space wavelength (in mm). For 3456MHz, $\lambda_o = 300/3.456$mm

$\lambda_o = 86.8$mm

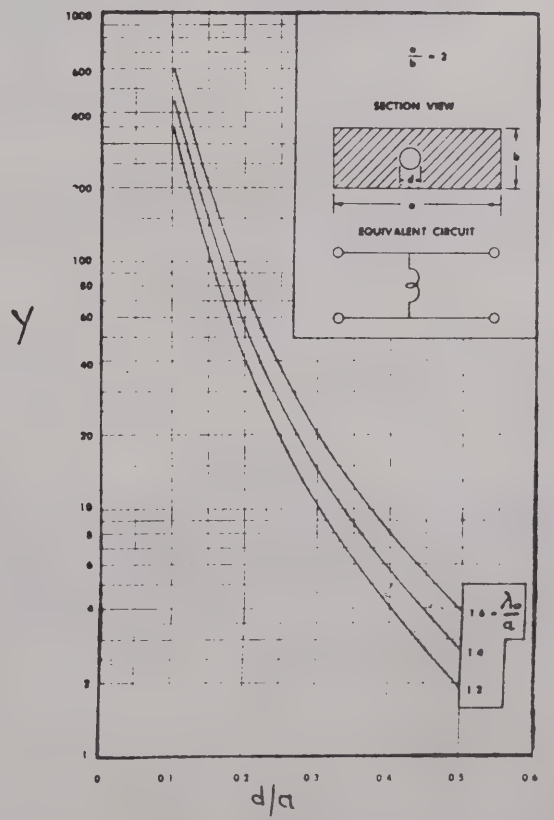
And so $\lambda_o/a = 86.8/60.25$

ie $\lambda_o/a = 1.44$

Graph for R=2.25



Graph for R=2.00



Then read off for each value of Y the corresponding iris hole size. The size is in terms of d/a, and so to obtain the actual diameter, d, this value must be multiplied by a. Use the set of curves most appropriate for the aspect ratio. Use the line from the set most appropriate for the ratio λ_g/a.

For the example given, using the curves for R=2.00

$d_{01}/a = 0.370$
$d_{12}/a = 0.190$
$d_{23}/a = 0.190$
$d_{34}/a = 0.370$

Also, for the example, using curves for R=2.25

$d_{01}/a = 0.360$
$d_{12}/a = 0.180$
$d_{23}/a = 0.180$
$d_{34}/a = 0.360$

Now for WG11, R=2.11, so to a first approximation, use the avarage of the two sets of values (R=2.125). The values are not that different that more precise interpolation is needed. Thus use values of 0.365 and 0.185. For WG11, a=60.25 and so the iris hole sizes calculate to be:

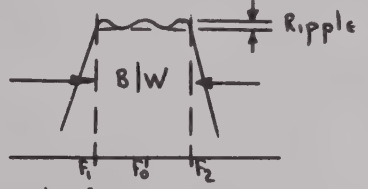
$d_{01} = 23.2mm$
$d_{12} = 11.1mm$
$d_{23} = 11.1mm$
$d_{34} = 23.2mm$

This completes the design of the filter, which is summerised below. Hopefully the information will enable filters to be designed for any microwave band in any waveguide. For anyone requiring more information, contact either the editors or G3JVL. We hope to add to this info in the future, but a complete treatment of the subject would take a number of issues!

Further to last month's information, here are some more notes on waveguide filters.

1. Correction

The bandwidth quoted in (2) last month should have been what is known as the ripple bandwidth (not the 3dB bandwidth). See diagram below. The calculations are otherwise correct, but in general (for values of passband ripple less than 3dB) the 3dB bandwidth obtained will be larger.



2. Normalised element values

The values given in (6) last month for g_0 to g_4 were for a Tchebyscheff response with 0.2dB of ripple in the passband. Here are some more sets of values for other amounts of passband ripple:

0.1dB Passband Ripple	0.5dB Passband Ripple
$g_0 = 1.0000$	$g_0 = 1.0000$
$g_1 = 1.0315$	$g_1 = 1.5963$
$g_2 = 1.1474$	$g_2 = 1.0967$
$g_3 = 1.0315$	$g_3 = 1.5963$
$g_4 = 1.0000$	$g_4 = 1.0000$

3. Holes in Iris Plates

As an alternative to the graphs given in (10) last month, the equations do exist to calculate the size of iris hole required for a given value of Y.
Firstly, for a given Y, calculate what is known as the magnetic polarizability, M.

$$M_{01} = \frac{ab\lambda_{go}}{4\pi Y_{01}}$$

$$M_{12} = \frac{ab\lambda_{go}}{4\pi Y_{12}} \quad \text{etc.}$$

For a small circular hole in a thin plate the diameter required is simply:

$$d_{01} = (6M_{01})^{1/3}$$
$$d_{12} = (6M_{12})^{1/3} \quad \text{etc.}$$

However, a practical iris is not very thin and often not small. The value of M can be corrected to take into account the size and thickness. Unfortunately an exact correction is difficult because the equations are not simple - to make the correction it is necessary to know the final hole diameter. But a very close approximation can be obtained if the value of d obtained above is used. So calculate corrected values of M:

$$M'_{01} = M_{01}\left(1-2.912\frac{d_{01}^2}{\lambda_0^2}\right)10^{-\left\{\frac{1.6t}{d_{01}}\sqrt{1-2.912\frac{d_{01}^2}{\lambda_0^2}}\right\}}$$

(t = thickness)

$$M'_{12} = M_{12}\left(1-2.912\frac{d_{12}^2}{\lambda_0^2}\right)\times 10^{+\left\{\frac{1.6t}{d_{01}}\sqrt{1-2.912\frac{d_{01}^2}{\lambda_0^2}}\right\}}$$

etc.

These corrected values of M are then used to calculate corrected values of d:

$$d'_{01} = (6M'_{01})^{1/3}$$
$$d'_{12} = (6M'_{12})^{1/3} \quad \text{etc.}$$

The thickness of the iris plate is the value t in the correction equation, and it should be in the same unit as d, ie in mm.
The correction for M could now be repeated, but using more accurate value now available for d. The original value of M should still be used. In general, though, o correction will be sufficient.

Example:
For $Y_{01} = 7.9$ (see example last month) then

$$M_{01} = \frac{60.25\times 28.50\times 124.95}{4\times\pi\times 7.9} = 2161.2$$

$$d_{01} = (6\times 2161.2)^{1/3} = 23.5mm$$

Then calculate (1.0mm iris) $M'_{01} = 2302.6$

and so $d'_{01} = 24.0mm$

The value is similar to that obtained graphically last month - 23.2mm. Note that the above value includes a thickness correction (1.0mm thick iris plate) while th graphs do not, so exact agreement was not expected.

Section 12 Antennas

12.1 G3JVL Flyswatter Antenna

G3JVL is back on 10GHz again after the winter gales took the reflector plate off his flyswatter system. The new reflector plate is slightly larger than the old one, but made from a wire mesh that is very much more transparent to the wind. He has done some experiments with several meshes, and found that galvanised wire mesh with half inch square holes has a very low wind resistance. For use on 10GHz it needs two thin enammelled wires threading along the mesh in each hole along the direction of polarization to make it suitably reflective.

12.2 Dishes

12.2.1 Use of Sperical Reflectors as Approximations to Paraboli -G3YCF

For reasons of cost and ease of manufacture, many people have considered using spherical reflectors instead of parabolic ones for dishes. This is fine for dishes on the UHF bands, where useful gains can be achieved before the difference in shape reduces the antennas efficiency, but on the microwave bands, this is of little value, as these gains can mostly be achieved with horn aerials anyway.

The difference in profile depends on the f/d and the size in terms of the wavelength, and above a certain gain, the aperture efficiency of the dish decreases fairly rapidly with frequency.

The table below gives some values for the maximum useful gain that can be achieved from a spherical reflector - it is the gain at which the spherical one is 1dB down on the correct parabolic shape of the same diameter.

f/d	G max
.32	17dBi
.375	19dBi
.48	27dBi
.60	32dBi

12.3 Baluns

12.3.1 A Balun for use with 0.141" semi-rigid coax - G3JVL/G3YCF

This is a design by G3JVL for a 4:1 balun to match a balanced 200 ohm load to 50 ohm coax, with a bandwidth of about 10 - 15 % of the centre frequency. In principle it will work at any frequency, but physical constraints limit it to the range approx 400MHz to 10GHz.

Details of the design are given in Fig 1. It is made by cutting two slots along the opposite sides of the coax outer with a junior hacksaw. The cable can be held in a vice and bent slightly while the cuts are made, as in Fig 2. Try to avoid cutting into the dielectric too much.

The two 'leaves' formed by the coax outer form a twin wire transmission line which is a quarter wave long, and short circuited at one end. This quarter wave resonator is excited by connecting the coax inner conducter to the end of one of the leaves.

The two leaves should be kept in contact with the ptfe dielectric, and not bent apart at all. At lower frequencies, where the leaves are quite long, it may be necessary to hold them together with a few turns of insulating tape, or perhaps plastic washers pushed over them. The length of these cuts must be an electrical quarter wave, and since the space between the two leaves contains ptfe, and the space outside them is air, the effective dielectric constant gives a velocity factor of about 0.86, so the cut length, L, is 0.86 times the free space quarter wavelength. If there is a significant air gap between the leaves and the ptfe, then the velocity factor will be slightly higher. The balanced output is then taken from the ends of the two leaves.

Its operation can be checked by soldering a small 200 ohm resistor across the ends, and measuring the swr at the 50 ohm input. The result is shown in fig 3, the 200 ohm resistor probably has a significant reactive component at this frequency, and explains why the swr in not perfect at the centre frequency. This balun is used to feed the slot in the Alford slot aerial.

Briefly, the balun works by taking the voltage on the unbalanced 50 ohm line and producing two output voltages relative to earth (the cable outer), which are equal to the input voltage, but are 180 degrees out of phase with each other, fig 4. The balanced load is connected between these two outputs, and sees the difference between them, which is twice the 50 ohm voltage, hence the 4:1 step up in impedance.

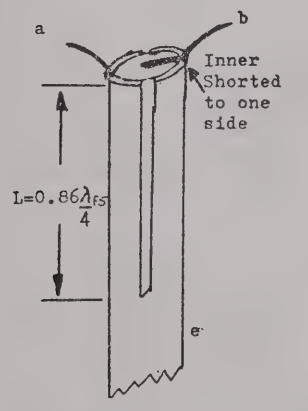


Figure 1

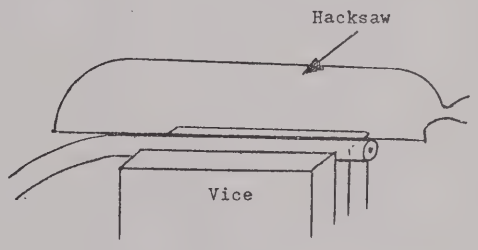


Figure 2

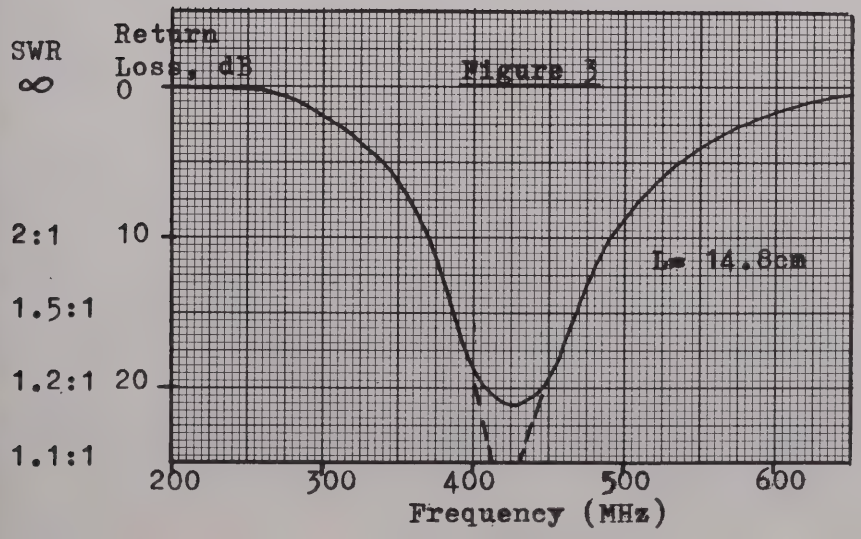


Figure 3

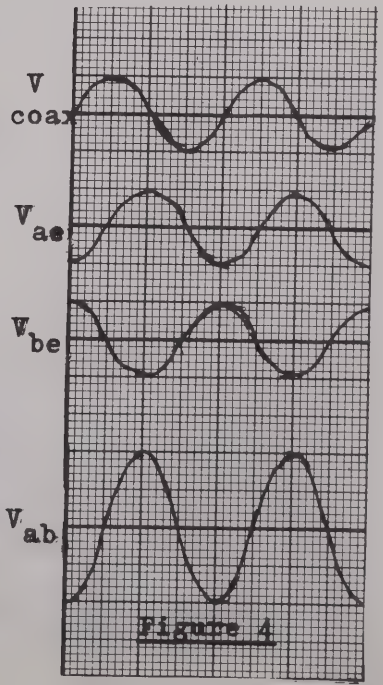


Figure 4

12.4 Alford Slot Antennas

12.4.1 page 1 Theory of Operation - G3JVL/G3YGF

Introduction

The Alford Slot antenna, which has been developed for 1.3GHz by G3JVL, is an easy means of obtaining an omni-directional radiation pattern with horizontal polarisation. The antenna has a gain which depends principally upon its length and is typically 5 to 9 dBi. This is a better performance than other simple omni-directional antennae commonly used such as halos or whips.

It is particularly suitable for a beacon or repeater antenna where an omni pattern is required with as high a gain as possible. In this application it is possible to stack two such antennae end to end (as used at the beacon GB3IOW) and nearly double the gain. With higher path losses on 23cm compared to 2m and 70cm the extra gain makes it particularly useful as a mobile antenna.

The bandwidth is sufficient to cover all of the band so that it would be suitable for any modes including TV. The circularity is very good (ratio of max to min gain) being typically 1 dB. This type of antenna has also been used on other bands successfully - G3JVL has used it on 2m, 70cm and 13cm. For further details contact Mike Walters G3JVL, or Julian Gannaway G3YGF, or the RSGB Microwave Committee.

Description

The antenna consists of a length of slotted tubing as shown in figures 1 and 2. By suitable design the feedpoint can be either in the centre (fig 1) when the slot is shorted at both ends, or at one end when only the other end is shorted (fig 2).

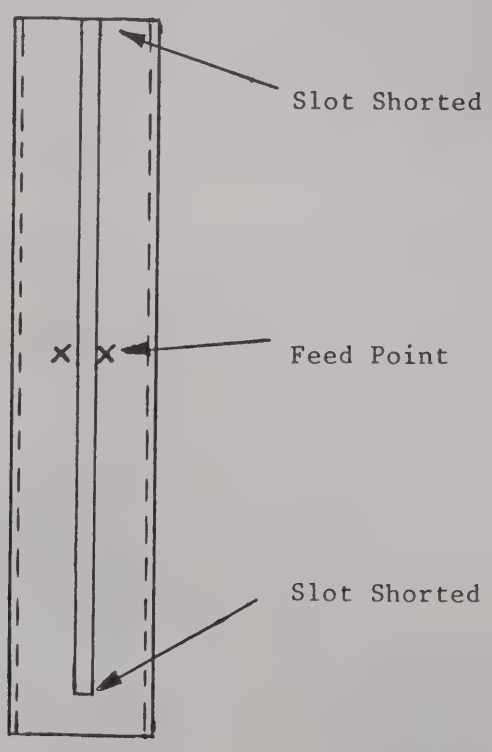


Figure 1

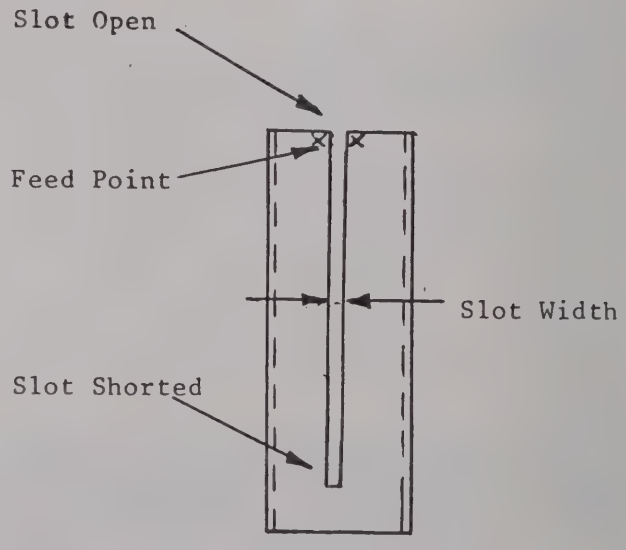


Figure 2

Operation

Slot antennae are not new - a vertical half wave slot is equivalent to a horizontal half wave dipole and produces horizontal polarisation. The novel feature of the Alford is that by making the wave travel up the slot faster than light it is possible to obtain a dipole type field distribution over its length which is many tim longer than the free space half wavelength value. The net gain is similar to that obtained by feeding several dipoles in phase but is obtained without the need for a complicated phasing harness. The gain obtained is directly proportional to the length of the slot in free space half wavelengths.

The idea that waves are travelling faster than light would at first sight seem impossible but in fact it is only a standing wave pattern that appears to travel at this speed; the actual wave travels at a lower velocity than light. The velocity of the wave along the slot varies as shown in figure 7. This increase is very similar to that which occurs in waveguide near cut off frequency (F_{co}).

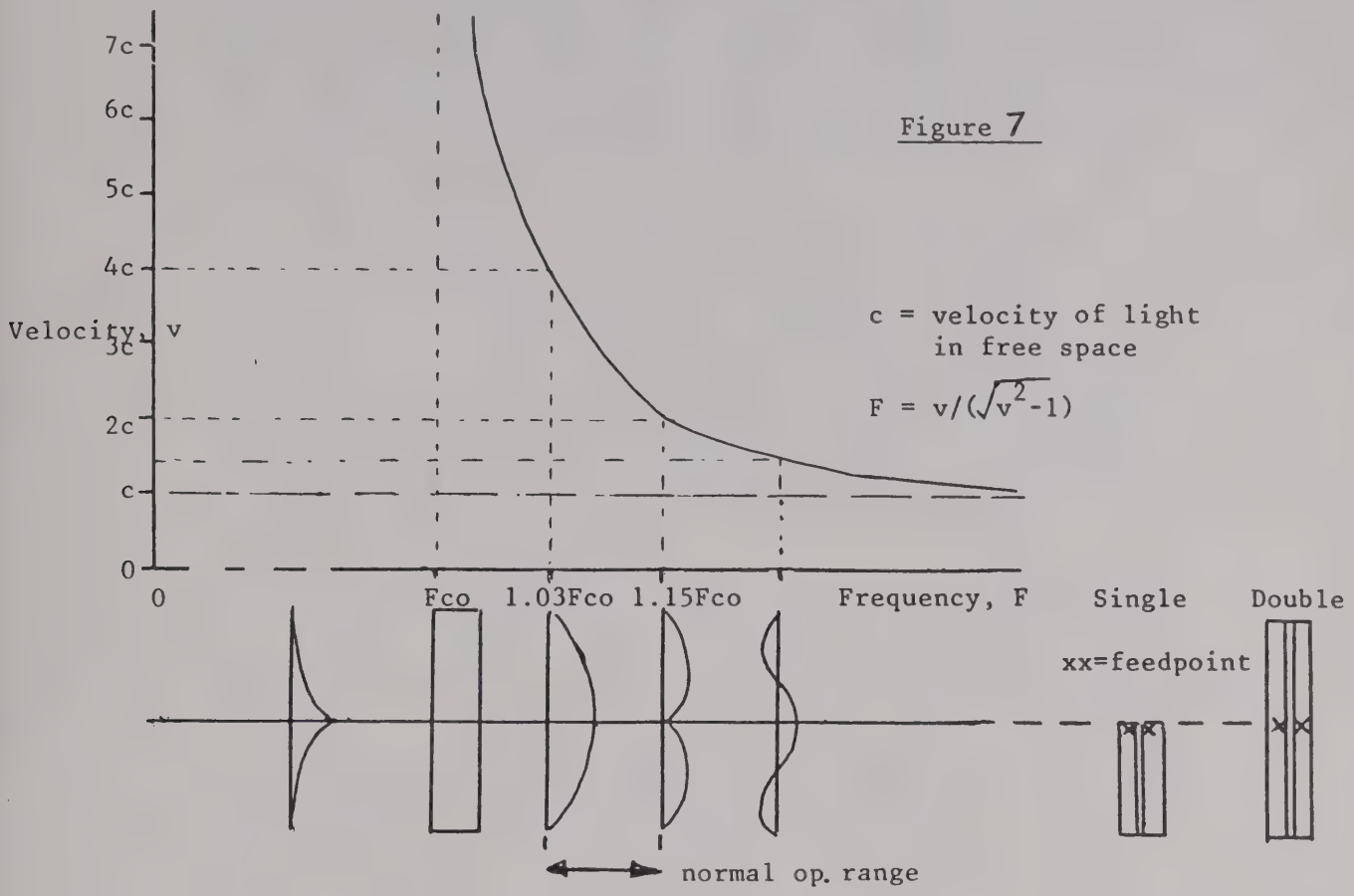


Figure 7

The slot behaves like a transmission line shunted by inductive loops (the solid cylinder is equivalent to an infinite line of closely spaced loops). Cut off occurs when the shunt inductance resonates with the capacitance of the slot. Below the cut off frequency waves cannot propagate at all. At the cut off frequency the velocity (and hence wavelength) is infinite. Above the cut off frequency the wavelength eventually decreases to the free space value.

In principal any velocity factor could be used but the higher the velocity factor (longer the slot) the more critical the dimensions. Velocity factors greater than about 10 are impractical for this reason and the normal operating range is around 5 to 15% above cut off ie with velocity factors of 2 to 5.

The dimensions are, to a certain extent interdependant. The velocity factor will be increased by decreasing the tube diameter, or by increasing the slot width. The wall thickness also has an effect since it determines the capacitance across the slot so that a thinner walled material will also increase the velocity factor. Thus if a slightly smaller diameter tube was chosen than one of the designs then this could be compensated for by using a slightly narrower slot so that the same velocity factor is achieved. Alternatively the length of the slot could be decreased. The antenna would then operate with a lower velocity factor, but this would give a lower gain.

It is important that the operation is checked, particularly if any of the original design parameters are changed. This may be done by feeding the antenna with a signal at various frequencies and looking at the voltage distribution using a power meter, detector or analyser with a small probe to pick up the radiated signal (see figure 8). The probe should be held close to the tube, but not directly in front of the slot (hold it 20 or 30 degrees round from the edge) and moved along its length. The diode current meter described in the microwave newsletter (08/81) would be suitable for this purpose.

Bibliography

1. Radio Communication: August 1981 page 732 (RSGB).
2. Microwave Newsletter: 02/82 and 08/81 (RSGB).

Section 13 Bearings

National Grid references provide a quick and simple means of calculating bearings within the UK (see last month). Unfortunately this system of maps is not available on the continent and so other longer or less accurate methods have to be used. However, the equations to convert between Lat/Long and NGR do exist so once the NGR of a foreign station has been calculated quick bearing calculations to it from any site can be made in the usual way.

The equations for this conversion were published by HMSO in 1950 in a booklet entitled "Constants, Formulae and Methods used in the Transverse Mercator Projection". These complete equations are somewhat complicated but for an accuracy of only 100m, which is probably all that is needed for microwave work, they can be simplified considerably. In doing this, G4KNZ "discovered" one or two interesting facts about the grid and its history and some of these which are probably not generally known have been included.

History of Triangulation

It is interesting to read an account of the triangulation of Britain which was necessary to produce the OS maps we are now familiar with. A lot of surveying work was done in the late 1700's and 1800's but this resulted in a large number of semi-independent triangulations of various portions of the country which did not fit together as a whole. Thus in 1936 what is known as PRIMARY TRIANGULATION began to accurately link the whole of England, Wales and Scotland and the network was finished in 1953. This involved surveying over paths of 40 to 60km (nearer 20km in eastern counties) to establish a large network of interlinked points. Observations were made a number of times to reduce errors, and readings normally made from nightfall to 3am! The results of triangulation were adjusted so that all triangles closed to exactly 180 degrees and side lengths could be computed through any route without any variation in the result. The average error in triangular misclosure is quoted as being 1.2 seconds of arc. Many of the points used were well known microwave sites including Butser Hill, Inkpen Beacon (Walbury) and others. From these primary stations, secondary and tertiary observations were made to fill in the areas between.

Transverse Mercator Projection

Many different types of projection of the curved surface of the earth onto a flat sheet are possible when trying to produce a map of the UK. One fundamental problem with any projection is that distances on the projection do not remain true to scale over a large area. Also, the scale error does not, in general remain equal in all directions at a given point (ie the scale in the NS direction is different to that in the EW direction). This results in direction errors which cannot be allowed for (in accurate work) by simple means.

However, one type of projection, known as the Transverse Mercator Projection, has the useful property that at any given point the scale is the same in all directions and it is for this reason that it is used

for the maps of Great Britain. In the simple transverse mercator projection, the surface of the spheroid chosen to represent the earth is projected onto a cylinder which touches the spheroid along a chosen line of longitude (meridian), and is then unwrapped. For the UK this chosen meridian (known as the central meridian) is the line of longitude 2 degrees West. The scale is then correct along this meridian and slowly increses with distance E and W from it, but at every point the scale is the same in all directions.

In the national projection the scale has been made too small on the central meridian by a factor of approximately 0.9996. The projection then becomes true to scale about 2/3 of the way between it and the edges of the projection. On the edges the scale has increased to about 1.0004 and the maximum error, which would otherwise have been 1.0008 at the edges, has been reduced.

Local Scale Factor

Thus when calculating distances accurately using NGR's the following scale factors should be applied:

National Grid Easting		Scale Factor, F
400	400	0.99960
350	450	0.99963
300	500	0.99972
250	550	0.99988
220	580	1.00000
200	600	1.00009
180	620	1.00020
160	640	1.00031
140	660	1.00043

True distance = distance on projection/F

As an approximation, for a path not due N or S, the mid point value of eastings should be used in determining the scale factor. More accurate formulae for the correct value of F are available but to put this change of scale into perspective, on a 250km path the error is 100m or less - hardly worth considering! For anyone calculating NGR's from Lat/Long, however, it cannot be neglected and is included in the constants. Also, if the grid is extended much E or W of the UK, the errors begin to become significant.

Calculating Bearings - G3YGF

The National Grid is a rectangular grid drawn on a transverse Mercator projection map of the UK, with the prime meridian at 2 degrees west. This method of calculation assumes that the paths are straight lines on this map, which is not exactly correct, as this map is a flat representation of a spherical earth. In practice this method is very accurate over paths of up to 1000km.

Most people will be familiar with the 6 figure reference where a point on the map is specified by the two letter reference for the 100 km square, and the number of km East and North of the bottom left hand corner of that square. Thus SZ710992 represents a point 471.0 km East and 99.2 km North of the grid origin.

The 100 km squares can also be specified by their number of hundred km from this origin, giving an eight figure reference. Thus 47100992 is equivalent to SZ710992. This latter method is far more useful for calculating bearings as Eastings and Northings can be obtained directly by subtraction of the references, rather than having to work out where "SZ" square is relative to your own, so it is far less prone to errors. These numbers of hundred km can be found at each corner of an OS map, where they are in small print by the numbers of km.

The bearing and distance are calculated by the simple geometry of a right-angled triangle, Figure 1. The difference between the northerly and easterly coordinates of each station are required.

If the stations have coordinates He, Hn, De, Dn (Home east, north, and distant east, north) then the eastings (ΔE) are De-He and the northings (ΔN) are Dn-Hn then:

$$X = \arctan \frac{\Delta N}{|\Delta E|}$$

where $|\ |$ means take the positive value of ΔE, even if it is negative.

Then if ΔE is positive,	Bearing TO	= 90 - X
	Bearing FROM	= 270 - X
and if ΔE is negative,	Bearing TO	= 270 + X
	Bearing FROM	= 90 + X

If ΔE is zero then the other station will be due north or south of you, (ΔN +ve or -ve), and the expression for X will become infinite.

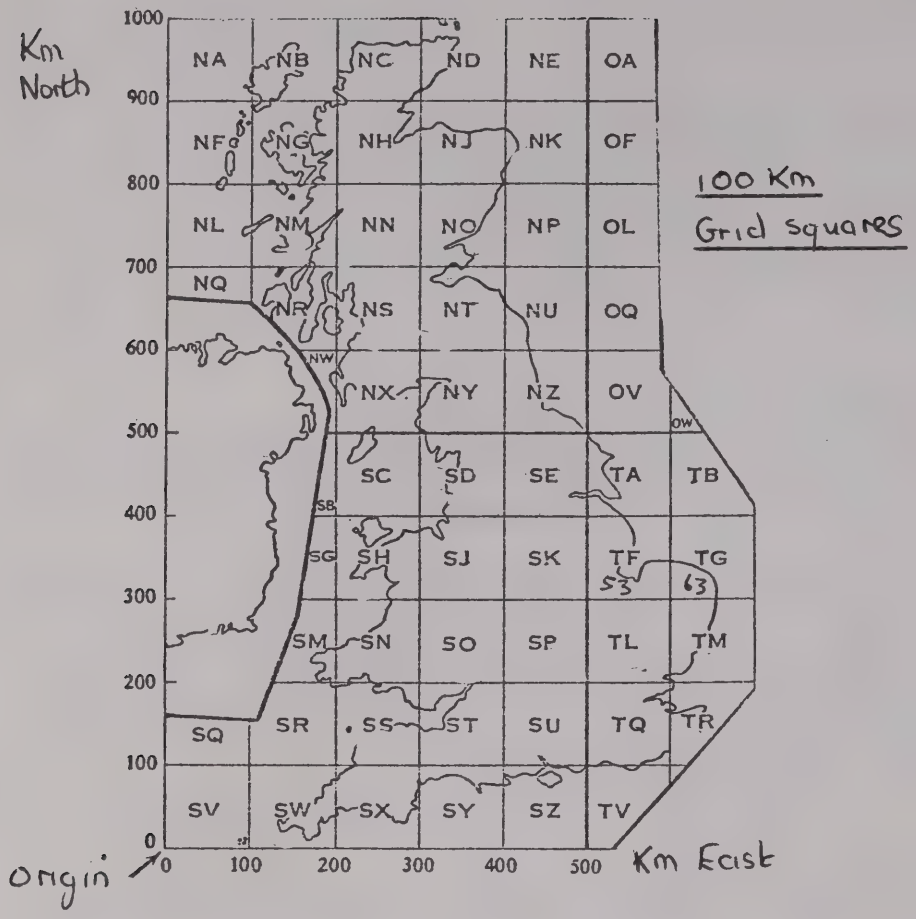
Also:

$$\text{Distance} = \sqrt{\Delta E^2 + \Delta N^2}$$

This method gives bearings which are exactly 180 degrees apart for each station, and relative to grid north, which differs from true north by up to 3 degrees, depending on the location. The exact difference is given on each OS map, and is typically 3°W on the east coast of England, 0° on the longitude 2°W, and 1°E in Wales.

The difference for a station in the UK can be found approximately by taking the difference between the longitude of the station and that of the 2 degree West line (degrees West are positive, degrees East negative), but this can be up to about 0.5 degrees in error.

Magnetic north is at present about 7°W of true north, but this also varies with locality, being 9°W in Wales and 6°W on the east coast. The actual value should be taken from the OS map.



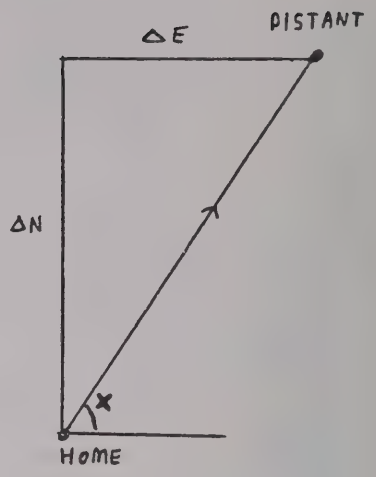


Figure 1

13.1.3

Cross Channel Bearing calculations - G3JHM

One problem with contacts across to the continent is that of calculating the bearings. Whilst the continental stations may know their QRA's it is a very long process to convert the QRA back to lat and long and then calculate the bearings.

Bearings in the UK can be calculated very easily from NGR coordinates but the system does not officially exist outside the UK. However if the lat and long is known it is possible to calculate the NGR and in the case of French stations all this means is that the northings become negative. This is no problem as long as the sign is used consistently throughout the calculation. The bearings calculated thus will be quite accurate enough for most purposes.

G3JHM has compiled the following list of NGR's for a number of commonly used sites in France.

Site / Location	km East	km North	QRA
Octoville, 10km N Le Havre	553.8	-32.6	AJ31j
1km NE Fecamp	572.8	-14.6	AJ14h
Etretat	558.1	-17.7	AJ22h
Cap de la Heve Le Havre	553.8	-32.6	AJ31e
Mt Casigny, 5km SW Deauville	548.5	-58.7	AJ51h
1km E Arromanches	500.2	-59.9	ZJ57j
Colleville, 5km W Pont Bessin	480.9	-58.3	ZJ56b
Mt Pincon, 60km SW Caen	501.0	-102.4	ZI07f
La Pernelle, 6km SW Barfleur	449.4	-30.9	ZJ34a
Jobourg, 25km W Cherbourg	407.2	-21.9	ZJ22j
Gatteville, 3km N Barfleur	452.7	-21.9	ZJ24b
GB3ALD, 1km NE St Anne	386.5	-20.0	YJ30h

13.1.4

NGR Programme

The following program was written by Cliff Barber, G4BGP for use on a pocket computer such as the SHARP PC-1211 or the TANDY TRS-80. It was written as a solution to calculating bearings in the field, since bits of paper etc would often end up blowing away in the wind. The above pocket computers retain the program when switched off, which can be recalled every time such a calculation is necessary.

The programming language is Basic, and could be used with most Basic interpreters with little or no modification. A listing of the program is below:

```
10  "A":INPUT"MY NGR LETTERS ?",B$,"FIGS ?",C:GOSUB 100:K=E:L=F:BEEP 2
15  INPUT"HIS NGR LETTERS ?",B$,"FIGS ?",C:GOSUB 110
17  G=E-K:H=F-L:I=√(G*G+H*H)
20  IF H<>0 GOTO 25
21  IF G>0 LET J=90:GOTO 80
22  IF G<0 LET J=270:GOTO 80
23  PRINT"YOU I/P OWN NGR IN ERROR":GOTO 15
25  J=ATN(G/H)
30  IF H<0 LET J=J+180:GOTO 80
40  IF G<0 LET J=J+360
80  BEEP 2:PRINT USING"####.#";I;"KM AT ";J;"DEGREES":GOTO 15
90  END
100 A$(21)="SW",A$(22)="SR",A$(23)="SM",A$(31)="SX",A$(32)="SS",A$(33)="SN"
101 A$(34)="SH",A$(35)="SC",A$(36)="NX",A$(41)="SY",A$(42)="ST",A$(43)="SO"
102 A$(44)="SJ",A$(45)="SD",A$(46)="NY",A$(47)="NT",A$(51)="SZ",A$(52)="SU"
103 A$(53)="SP",A$(54)="SK",A$(55)="SE",A$(56)="NZ",A$(61)="TV",A$(62)="TQ"
104 A$(63)="TL",A$(64)="TF",A$(65)="TA",A$(72)="TR",A$(73)="TM",A$(74)="TG"
110 FOR D=21 TO 74:IF B$=A$(D) THEN 120
115 NEXT D: PRINT"ERROR IN NGR LETTERS": GOTO 15
120 E=INT(D/10)*100+INT(C/1000)/10:F=100(D-10*INT(D/10))+C/10-100*INT
    (C/1000)
140 RETURN
```

The user is asked to enter his NGR letters and NGR figures separately. A subroutine in the program takes the NGR letters and compares them with a list of selected letter groups.

Each letter group is assigned a value which is related to the eastings and northings of the NGR letters code. (eg TL=63). This value is used along with the NGR figures entry to calculate eastings and northings to the nearest 100 metres as defined by the 6 figure reference. (Note: this value is only related to the actual NGR, being in fact 100km different to the true NGR in both eastings and northings. This makes no difference to the calculation and was necessary due to the way the two computers mentioned above organise their memory.)

The user is then asked for the NGR of the other station. The subroutine again sorts the data and calculates the new eastings and northings. Finally the program calculates range using Pythagoras, and the angle, sorting out the relevant quadrant. About a minute is needed to do all this, and a typical result: "94KM AT 316.4 DEGREES".

In its present form the program does not include all of Scotland, but could be easily modified to do so. The print statement on line 115 is intended to tell the user that the NGR letters he entered were not stored in the program.

For use on one of the above mentioned pocket computers, having loaded the program type in "CLEAR", terminated by the ENTER key, followed by "RUN 10", teminated by the ENTER key. Obviously if the program is adapted for other machines these commands may be different.

Note that the bearings calculated are grid bearings, and a correction of approx 7 degrees is required when transfering it to a compass (add 7 to obtain magnetic).

13.1.5

NOTES ON THE N.G.R. PROGRAMME BY G4BGP. - G3HWR

The above programme can be simplified and improved to include the whole area covered by the National Grid.

The simplification depends on the fact that all sensible computers represent the alphabet by a sequence of 26 binary numbers, but this sequence varies from machine to machine though the increment is constant for any machine. This increment is "B" - "A" and, for example ("X" - "A")/("B" - "A") = 23 for all machines.

The following notes are intended to help anyone adapt the programme section to varieties of BASIC. The same line numbers are used as in G4BGP's listing, so the material that is unchanged is not repeated here.

Line 100 determines the increments for the first and second letters of the N.G.R. respectively, it replaces the look-up table.

Line 110 obtains the integer corresponding to the first letter, i.e. 0 to 25 then, because "I" is not used, the value of "H" is fiddled to be consistent with the others. Line 112 evaluates the 500 km. steps in E and F using the following transformation:

H	J
N	O
S	T

8	9
13	14
18	19

0	500
0	500
0	500

&

1000	1000
500	500
0	0

Line 114 obtains the integer corresponding to the second letter, then reduces this value by one for all letters after "I" which is not valid. Line 116, which can be ommitted, guards against this erroneous use of "I".

Line 119 increases E and F by multiples of 100 using the following transformation:

A	B	C	D	E
F	G	H	J	K
L	M	N	O	P
Q	R	S	T	U
V	W	X	Y	Z

0	1	2	3	4
5	6	7	8	9
10	11	12	13	14
15	16	17	18	19
20	21	22	23	24

4
3
2
1
0

&

0	1	2	3	4
0	1	2	3	4
0	1	2	3	4
0	1	2	3	4
0	1	2	3	4

Lines 120 adds the remainder of the reference and corresponds to the same line in the original listing. If your machine prefers one long formula, then lines 112, 119 and 120 can be combined.

The power of this method is that where the look-up table threatens to be very large, e.g. for QTH (QRA) or GEOREF, it saves having to compromise by omitting squares that are seldom needed.

```
 23 PRINT "THAT IS NOT A VALID N.G.R."

100 LET X1 = "BA" - "AA" : LET X2 = "AB" - "AA"
110 LET N1 = INT((B$ - "AA")/X1) : IF N1 = 7 LET N1 = 8
112 LET E = 500 * (N1 - INT(N1/5) * 5 - 3) : LET F = 500 * (3 - INT(N1/5))
114 LET N2 = INT((B$ - "AA" - N1 * X1)/X2)
116 IF N2 = 8 THEN 23
118 IF N2 < 8 LET  N2 = N2 - 1
119 LET E = E + 100 * (N2 - 5 * INT(N2/5)) : LET F = F + 100 * (4 - INT(N2/5))
120 LET E = E + INT(C/1000)/10 : LET F = F + C/10 - 100 * INT(C/1000)
140 RETURN
```

13.2

Bearing Calculations using Lat and Long - G3YGF

This method gives exact values for the bearings and distances. The calculations can be done with a scientific pocket calculator and the bearings are given relative to true (geographic) North.
First the lat and long of both stations are converted into the form degrees and decimal fractions of a degree. The distant station's lat and long are referred to as DN and DW (Distant North, Distant West) and the home station's lat and long as HN and HW (Home North, Home West). Degrees North and West are positive, South and East negative.

DIFF represents the difference in longitude between the two stations, DW-HW and A represents the angle in degrees subtended at the centre of the earth by the two stations. The bearings and distance are then calculated as follows:

Distant	DN = **.****	DW = ***.****
Home	HN = **.****	HW = ***.****

DIFF = DW - HW

A = arccos(cos(DIFF)*cos(DN)*cos(HN)+sin(HN)*sin(DN))

Path Length = A*111.15 km

$$X = \arccos\left[\frac{(\sin(DN)-\cos(A)*\sin(HN))}{\sin(A)*\cos(HN)}\right]$$

$$Y = \arccos\left[\frac{(\sin(HN)-\cos(A)*\sin(DN))}{\sin(A)*\cos(DN)}\right]$$

If DW is greater than HW then: Bearing TO = 360-X
Bearing FROM = Y

Otherwise: Bearing TO = X
Bearing FROM = 360-Y

This method gives the exact great circle bearings relative to true North. It will be found that the bearings TO and FROM are not exactly 180 degrees apart. This is quite correct, and other methods such as NGR which give bearings exactly 180 degrees apart are only approximate.

13.3 Conversion Between Lat/Long and NGR

13.3.1 page1 Conversion of Lat/Long to NGR - G4KNZ

The following equations will give an accuracy of better than 25m over the UK, and when extended to cover Eire, Northern France, Belgium and Holland will give an accuracy still within 100m:

Firstly evaluate varable Y:

$$Y = (1.0-0.00667054*\sin^2(LAT))^{\frac{1}{2}}$$

Then, with LAT and LONG in degrees:

$$\begin{aligned}\text{Northings} = {} & -100.0000 \\ & +111.0793*(LAT-49) \\ & -31.9470*\sin(LAT-49)*\cos(LAT+49) \\ & +0.9710*(2-LONG)^2*\sin(LAT)*\cos(LAT)/Y\end{aligned}$$

$$\begin{aligned}\text{Eastings} = {} & +400.0000 \\ & +111.2651*(2-LONG)*\cos(LAT)/Y \\ & -0.0056*(2-LONG)^3*\cos^3(LAT)*(\tan^2(LAT)-1)\end{aligned}$$

where the result will be in km.
The full equations can be obtained from the reference earlier, but additional terms get progressively more complex. Note that degrees N and W are positive, and care must be taken to include a negative sign for points with an easterly longitude.

The true origin of the national grid is at 2 degrees W, 49 degrees N. For convenience, however, these points are designated to be 400km E and -100km N exactly so that the false origin where eastings and nothings are zero is approximately 49° 46'N, 7° 33'W. This means that over the UK all values of eastings and northings are positive. For calculations to the W and S of this false origin it will be found that eastings and northings become negative. This does not present any problem, provided the sign is preserved in all calculations.

In general, the errors in the above equations increase with distance from the true origin. There is no problem in calculating points up to 61 degrees N or so, and well below 49 degrees N. The main difficulty is trying to use the grid too far E or W when the above equations hold to 100m to approx 12°W and 8°E. However, if the variation of local scale factor (see earlier) is examined it is found that beyond the point where it is unity it begins to rise rapidly with distance further E°or°W. Distance calculations to these extremes of 12°W or 8°E should thus be treated with caution. Formulae for the calculation of F do exist and at these points the scale factor is around 1.005 (varies with lattitude). Again consult the reference if interested.

Alternative Equations

For quick calculations in the field the above equations are not the most convenient. An alternative is the following set of equations which were derived by G3JVL, G3YGF and G4KNZ from some equations for converting in the other direction by G8DIC.

Firstly evaluate $L = \arctan(\tan(LAT)/\cos(LONG-2))$

where LAT and LONG are measured in degrees, degrees N and W being positive.

Then evaluate B = arccos(sin(LAT)/sin(L))

Then if LONG>2 then Eastings = 400.00 + 111.55*B Km
Or if LONG<2 then Eastings = 400.00 - 111.55*B Km

Then evaluate X = (L-49)*(1.0+0.000079*(L-49))

And then calculate Northings = 111.166*X - 100.00 Km

These equations fit remarkably well around the S of the country and for Northern France (and are sufficient for use in calculating cross channel bearings). The accuracy will be well within 100m, which is more than enough. They are probably sufficient for paths to EI and ON/PA also. If in doubt, try both methods and compare!

13.3.2

LAT/LONG to NGR Conversion - G4KNZ

The following program converts lat & long to NGR using the equations in the 06/82 issue. It is written in BASIC and has been kept deliberately simple in the hope that it will run on most microcomputers. All the REM "comment" statements can be ommitted if desired.

```
1000 REM This program converts a lattitude/longitude
1010 REM to an NGR. Accuracy depends upon the distance
1020 REM from the true origin of the National Grid (at
1030 REM 2 degrees W, 49 degrees N), but is typically
1040 REM within 25m for most of the UK.
1050 REM
1060 REM Written by S J DAVIES, based on the Microwave
1070 REM Newsletter 06/82.
1080 PRINT "Lattitude (Degrees, Minutes N) ";
1090 INPUT D1,M1
1100 PRINT "Longitude (Degrees, Minutes W) ";
1110 INPUT D2,M2
1120 REM Convert to degrees and fraction of degrees.
1130 L1=D1+M1/60
1140 L2=D2+SGN(D2)*ABS(M2)/60
1150 REM Evaluate R, to convert degrees to radians.
1160 REM This is required because functions SIN & COS
1170 REM require the angles to be given in radians.
1180 R=1/57.29577951
1190 REM Evaluate SIN & COS of L1 which are used often
1200 S1=SIN(R*L1)
1210 C1=COS(R*L1)
1220 REM Evaluate variable Y.
1230 Y=SQR(1-.00667054*(S1^2))
1240 REM Calculate Northings in km.
1250 N=-100
1260 N=N+111.0793*(L1-49)
1270 N=N-31.947*SIN(R*(L1-49))*COS(R*(L1+49))
1280 N=N+.971*((2-L2)^2)*S1*C1/Y
1290 REM Calculate Eastings in km.
1300 E=400
1310 E=E+111.2651*(2-L2)*C1/Y
1320 E=E-0.0036*((2-L2)^3)*(C1^3)*(((S1/C1)^2)-1)
1330 REM Round off coordinates to 100m.
1340 N=INT(N*10+.5)/10
1350 E=INT(E*10+.5)/10
1360 REM Print out results.
1370 PRINT "Conversion to National Grid Reference:"
1380 PRINT "          Northings = ";N;" km"
1390 PRINT "          Eastings  = ";E;" km"
1400 END
```

Functions used are: SIN - sine
COS - cosine
SQR - square root
INT - truncate to an integer
ABS - absolute value of
SGN - sign of

Section 14 Radiation Hazards

14.1 Licencing

No doubt everyone is aware of the recent changes which were made to the licence schedule. On Friday 12th February a new schedule was published in the London Gazette which contained a number of changes made without consultation with the RSGB.

For microwave operators one serious change was the adoption of new methods of power measurement on bands above 1GHz. The power limit for these bands is specified as 30dBW (35dBW for ssb) eirp. It was also specified that at no time should the transmit power density exceed 10mW per square cm across the antenna aperture.

It is interesting to plot these restrictions on a graph for a band such as 1296MHz:

1. A line of 30dBW eirp can be drawn as a straight line on a graph of antenna gain vs power output. One must remain below this line.

2. For an antenna such as a dish it is possible to work out the power density across the aperture based on the power being distributed over its area. In practice the power is not distributed evenly over the area and for antennas such as yagis it is difficult to visualise an aperture but for a given gain we can obtain an approximate value of the power density for a given power input. This enables a second line of 10mW/cm^2 to be drawn, which one must remain above.

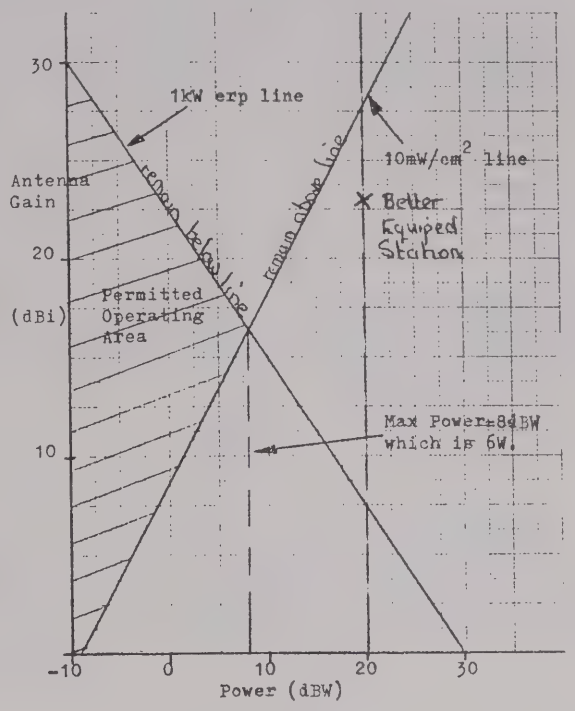
3. The old power output limit of 100W is shown for reference.

It can now be seen that these lines define a triangle within which it is possible to operate. Most significant is the maximum power it would be now possible to run, which is about 6W. An example of a better equiped 1296 MHz station as is commonly used on VHF NFD is marked on the graph - 100W to 23dBi and it can be seen it is a long way outside the operating triangle.

From the safety point of view, an eirp limit is in fact counter productive. For a given transmitter power then the bigger size of dish that was used, for example, then the larger the area over which the power is spread and so the lower the power density.

It is also possible to perform calculations on the power density some distance away from an antenna for a given power input. The calculations are rather complicated and difficult to do exactly. However, for 100W input to an antenna at 1296MHz it appears that the level of 10mW per square cm exists for quite a short distance. This varies with gain but the power has fallen below the level typically after several metres, even for high gain antennas (distance increases with gain).

Hence there is little argument for specifying a maximum power density provided that where the level of 10 mW/cm^2 is exceeded then people are not permitted access to the main beam. Note that off the main beam the density falls even more rapidly.



14.2 Measurements of Power Densities

G8SHF reports some measurements of power density 5 cm from an open X band w/g flange running 100mW as being 2-3 mW /sq cm. This was done with professional measuring equipment, and the values obtained agree well with calculated values. This illustrates that high power need not be a hazard, provided one keeps a sensible distance from the obvious regions of high power density.

14.2 Proposed RF Radiation Safety Limits

The National Radiation Protection Board have just produced a consultative document which reviews some of the existing international standards, and proposes new ones. The proposed safe levels are summarised below, and are the limits for continuous exposure of the general population, including children.

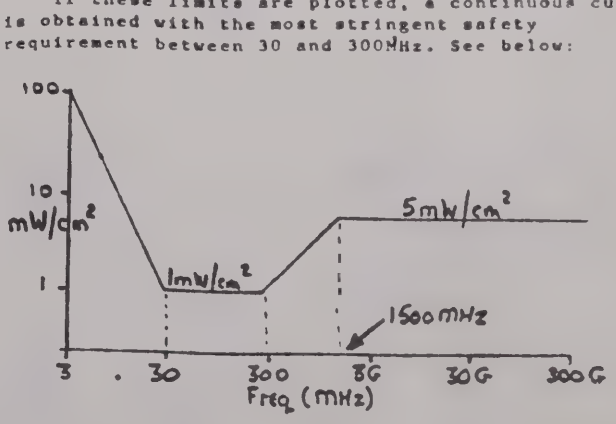
Frequency (MHz)	Power Density (mW/sqcm)
3 to 30	$900/F^2$
30 to 300	1
300 to 1500	F/300
1500 to 300000	5

From the microwave point of view, these are only a small change from the existing standards of 10mW/sq cm.

This is only a simplified summary of the proposals. There are certain conditions specified for the measurements, and under some circumstances the limits may be relaxed to some degree. The document contains a very good overview of the subject and is called "Proposals for the Health Protection of Workers and Members of the Public against the Dangers of Extra Low Frequency, Radio Frequency and Microwave Radiation: A Consultative Document". It is published by HMSO at a cost of £2.

The microwave committee has been considering this subject for some time, and will be giving the document careful attention.

If these limits are plotted, a continuous curve is obtained with the most stringent safety requirement between 30 and 300MHz. See below:

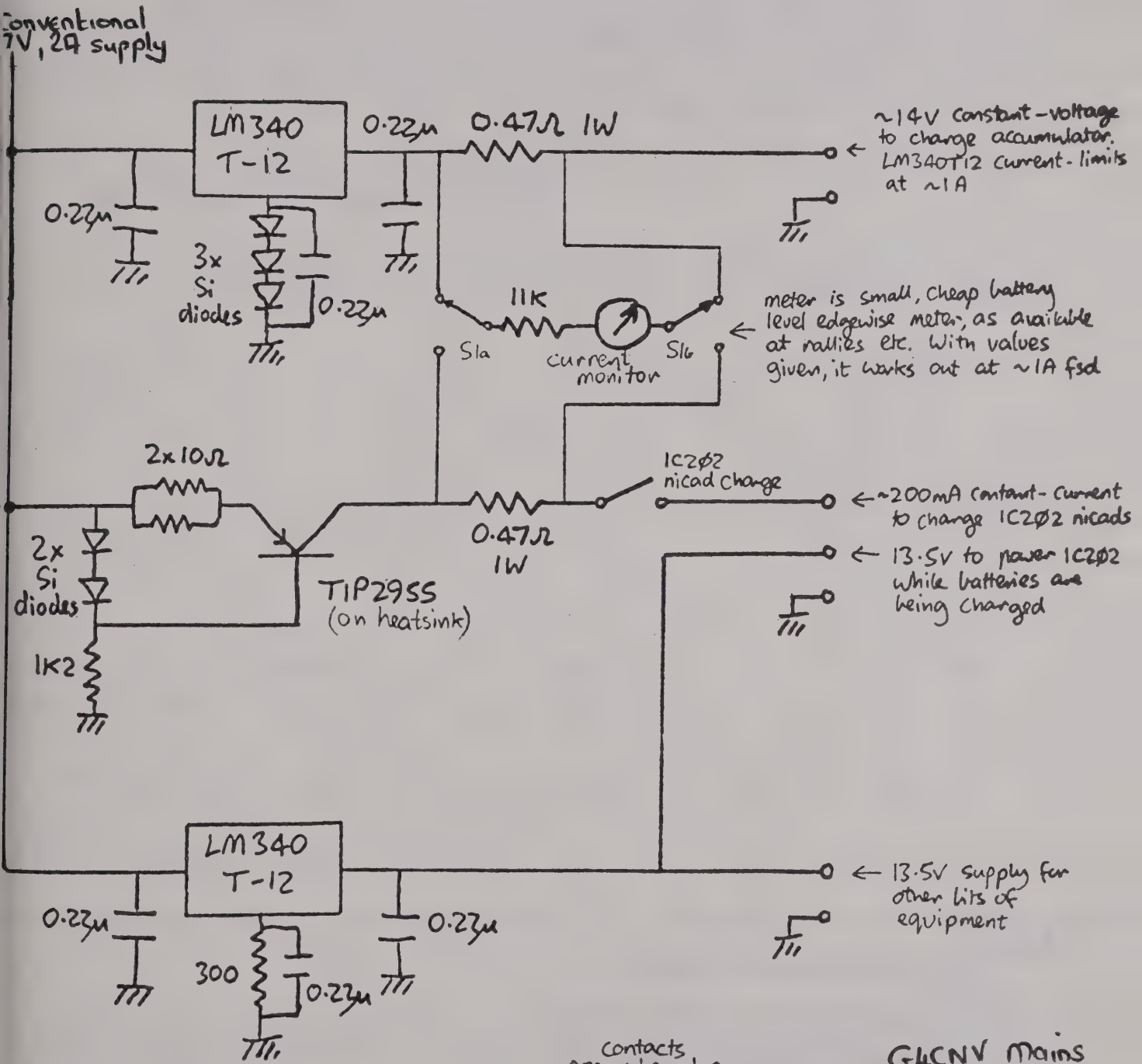


Section 15 Test Gear

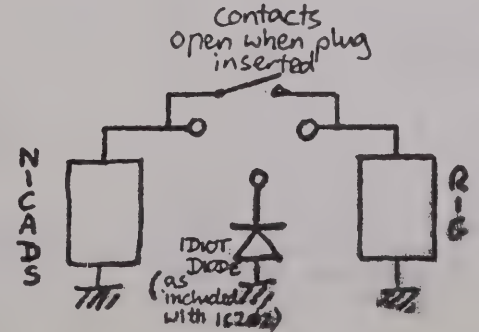
15.1 **Power Supplies and Voltage Regulators**

15.1.1 **Battery Charger/PSU - G4CNV**

G4CNV suggests a regulator circuit for charging IC2Ø2 nicads (at 200mA constant-current), charging a sealed lead-acid accumulator at 14V, current limited at 1A, and also providing a 13.5V 1A regulated supply for other gear. He has replaced the standard power socket on the IC2Ø2 by one from a TR10 2200 (available from Lowe's) which isolates the nicads when the plug is inserted, and means the nicads can be changed without removing them, while the rig is simultaneously powered by 13.5V from the psu. At the same time, the SO239 socket was replaced by a BNC. All this is perfectly neat and tidy and easy to carry out!



New IC2Ø2 Power socket. (Seen from outside of IC2Ø2) →



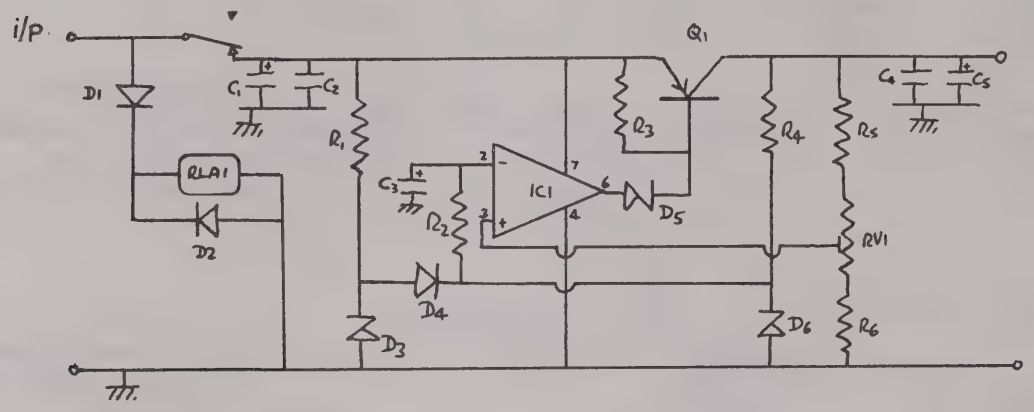
G4CNV Mains battery charger PSU

15.1.2 A Low Voltage-drop, Reverse supply Protected Regulator - G4CNV

For many pieces of equipment that are used portable, a well-regulated supply is essential to prevent drift, chirp and other problems, as the voltage from car batteries etc does vary somewhat according to state of charge, current drawn etc (typically 11.5 to 16 V). The following circuit was developed for use with the G3JVL 10 GHz transverter, and as well as making the oscillator well and truly insensitive to supply voltage variations, provides overvoltage and reverse supply protection. It will regulate down to a minimum voltage drop of approx 300 mV (governed by the V_{ce} sat of the transistor).

The relay used for reverse supply protection (RS 348-510) is rated at 1A, 28V. RV1 sets the regulated output voltage; something like 11.2V is suggested as a reasonable value that will guarantee a regulated output even at the lowest supply voltages. The BD132 transistor is bolted to a suitable heatsink (diecast box wall) with the usual insulating washer.

A pcb has been produced for the design, which should be available with the 384 / 378.666 MHz driver pcb's. The circuit diagram is reproduced below:



o/p voltage range

10.0 - 12.7 V nomi

R1	4K7	C1	10μF 25V	D 1,2,4	IN4149	IC1	741
R2	15K	C2	1nF disc	D 3,5	4.7V zener	Q1	BD132
R3	1K	C3	10μF 25V	D6	5.6V zener	RLA1	RS 348-510
R4	3K3	C4	1nF disc				
R5	18K	C5	100μF 25V				
R6	18K						
RV1	5K Cermet D						

This design is based upon a kit produced by Wood & Douglas

15.1.3

Protecting against reversing the polarity of supplies

One regularly hears of people blowing up their equipment by making wrong polarity connection to batteries, and this seems more common among microwave people simply because most of their operating is portable. Some of the methods used to prevent damage are shown in Fig 3. In (a) a series diode is fitted, which is satisfactory provided that the 0·5–0·7V drop across the diode is acceptable. In (b), a "crowbar" method, the voltage drop is avoided, but at the expense of blowing a fuse each time a mistake is made. If a power rectifier is used, then a high current fuse can be used which will have a lower resistance. A third technique which the writer has just come across (long after everyone else, of course) is shown in (c). The relay contacts simply remain open unless the polarity as determined by the relay series diode is appropriate.

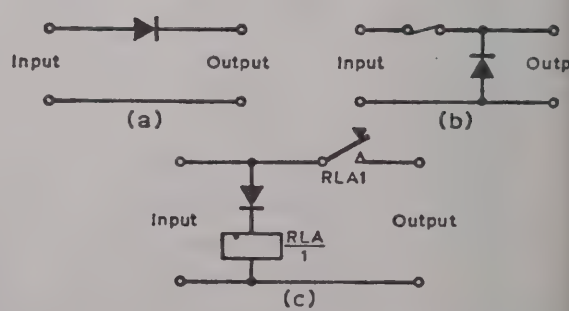


Fig 3. Circuits for protecting equipment against reversing the polarity of dc supplies

extracted from RadCom

15.2 Low Power 50 Ohm Loads G8KRD

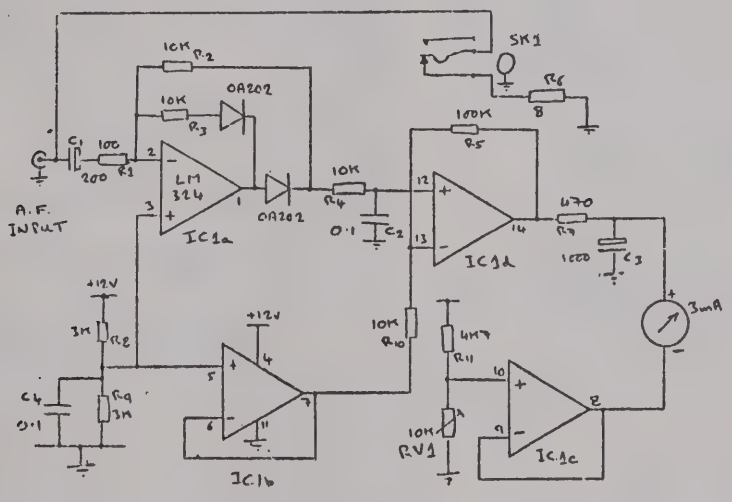
Low power 50 Ohm loads of quite reasonable VSWR can be made from unwanted crimp type BNC plugs. When a plug on the end of a piece of cable goes intermittant, it is usually not possible to reuse it, as spare crimp sleeves do not suit.

Choose a 50 Ohm film or composition resistor that just fits inside the sleeve to which the outer is normally crimped. Cut the resistor lead short at one end so it just fits into the pin, and solder it in. Insert the pin and resistor in from the rear of the plug, then solder the other wire end of the resistor to the sleeve, filling up the hole with solder.

Depending on the type of resistor used, the SWR can be better than 1.05 to 1 up to 1GHz.

15.3 Audio Voltmeter - G4ERP

From Richard Marshall, G4ERP, comes details of a simple audio voltmeter. With suitable calibration, or an attenuator substitition method, this could be used for measuring ground or sun noise, or measuring noise figure by the hot/cold technique :



"Introduction

The human ear, although a marvellous invention, is not particularly good at detecting small changes in audio levels. This is a distinct disadvantage when attempting to align a preamp or a receiver without the use of sophisticated Test equipment. The circuit described in this article is that of an audio meter which allows changes of less than 0.5dB to be observed. It has been in use for some time now, and has proved invaluable for aligning and comparing preamps, receivers etc and when testing aerials.

Circuit Description

The circuit of the meter is given in Fig 4. The basis of the unit is IC1 which is an LM324. This is a quad 741-type op-amp requiring only one power supply voltage, thus making it ideal for portable use.

The receiver audio output is applied to IC1a, which is connected as a precision rectifier. The raw d.c. is smoothed by the integrator network R_4 and C_2, and then amplified by IC1d. The output at pin 14 has a +6V d.c. offset, and this prevents the direct connection of a meter to 0V. To overcome this, IC1c generates on identical voltage so that the meter reads zero under no-signal conditions.

Resistor R is chosen to present the receiver with the correct impedance if the headphones should be removed from SK1. It should be of adequate power rating."

15.4 Attenuators

15.4.1

Design and construction of simple attenuators

The writer has often found the need for good attenuators capable of working at frequencies up to several hundred megahertz or higher in the course of microwave experiments. These are relatively easy to construct out of standard resistors, and can be put to a variety of uses. For example, attenuators at rf can be used as "pads" between interacting stages, eg varactor multipliers, or to follow noise or signal sources to bring their output impedance close to 50Ω. At i.f. they can be used for calibrating microwave attenuators, since their attenuation is fairly predictable at lower frequencies. They may also be used for calibrating S-meters etc, and as a reference for noise measurements, eg sun and ground noise.

Two simple configurations of symmetric attenuators are the "T" type and the "π" type, depicted in Fig 3. The design of these is covered by the formulas:

$$\text{Attenuation (dB)} = 20 \log \left(\frac{50 + R1}{50 - R1} \right)$$

when $$R2 = \frac{2{,}500 - R1^2}{2R1}$$

} "T" type

$$\text{Attenuation (dB)} = 20 \log \left(\frac{R1 + 50}{R1 - 50} \right)$$

when $$R2 = \frac{5{,}000R1}{R1^2 - 2{,}500}$$

} "π" type

Design values for some useful attenuation levels are given in Table 1. It will be seen that in most cases the values do not coincide with preferred resistor values. This restricts the levels of attenuation which can be obtained, and the best procedure seems to be to choose the nearest preferred values to R1 and R2. This usually results in an acceptably predictable performance, but where the approximations are too great it may be better to try the other configuration, or to use series or parallel combinations of resistors for R1 and R2.

Two methods of construction are illustrated in Fig 4. A "π" type is shown in (a) and "T" type in (b), but of course either version could be used in either design. The resistors should be of low inductance type, the common form of carbon film resistors being particularly suitable. Needless to say, lead lengths should be as short as possible. For higher power attenuators at lower frequencies, parallel combinations of 0·5W carbon resistors can be used to increase dissipation. A 10dB attenuator, built in this way to handle 10W, measured 9·5dB attenuation at 432MHz with low swr.

Provided that care is taken, these attenuators can be used up to 1-2GHz. The biggest error is likely to arise in the higher value attenuators, where stray coupling may reduce the attenuation below the expected value. For this reason it is better to use several low value stages in cascade when a high value of attenuation is required. □

Table 1. Design data for 50Ω "T" and "π" type attenuators

	"T" type		"π" type	
dB	R1	R2	R1	R2
1	2·9	433	870	5·8
2	5·7	215	436	11·6
3	8·6	142	292	17·6
4	11·3	105	221	23·9
5	14·0	82	178	30·4
6	16·6	67	150	37·4
7	19·1	56	131	44·8
8	21·5	47·3	116	53
9	23·8	40·6	105	62
10	26·0	35·1	96	71
12	30·0	26·8	84	93
14	33·4	20·8	75	120
16	36·3	16·3	69	154
18	38·8	12·8	64	196
20	40·9	10·0	61	248
25	44·7	5·6	56	443
30	46·9	3·2	53	790

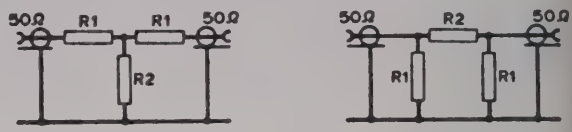
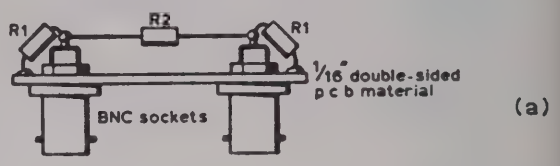


Fig 3. (a) "T" type attenuator; (b) "π"-type attenuator



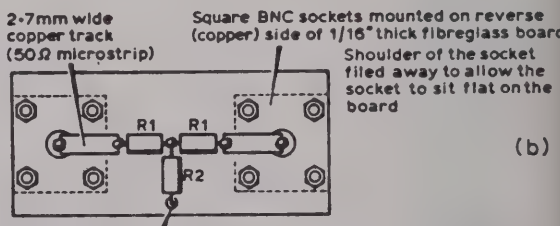

2·7mm wide copper track (50Ω microstrip)
Square BNC sockets mounted on reverse (copper) side of 1/16" thick fibreglass board.
Shoulder of the socket filed away to allow the socket to sit flat on the board
R1
R1
R2
(b)
Resistor lead passes through hole in the board and soldered to the other side

Fig 4. Methods of constructing attenuators: (a) method of construction below 500MHz; (b) method of construction above 500MHz

extracted from RadCom

15.4.2

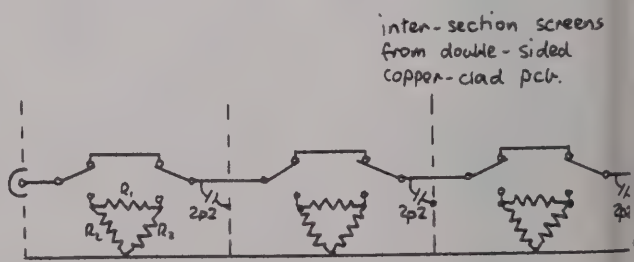
Switched Attenuators (G4CNV)

G4CNV has built a switched attenuator for use as a part of a noise-measurement system, to be used at 30MHz. By setting it up against an HP435B power meter, an accuracy of better than 0.05dB was obtained over a 60dB range. The reason why such an attenuator is such a useful part of a measurement system of this type is that it is far easier to build an accurate attenuator than an amplifier/detector capable of operating accurately over the same dynamic range. Essentially, the attenuator reduces the dynamic range required from the detector to 1dB, which can be met much more easily. The point should be made, however, that an attenuator that is to be used as part of a measurement system should be checked against good laboratory equipment.

Out of interest, the performance at other frequencies was also checked, and it was found that the characteristics had a resonance at about 90MHz (believed to be due to the switches used), and that the attenuation tended to increase at higher frequencies. This 'droop' could be considerably reduced by adding small capacitors (2.2pF subminiature ceramic disc, RS 125-496) in parallel with the switch sections. These capacitors and the inherent inductance of the attenuator elements form a sort of giant low-pass filter, with a cut-off frequency at about 1GHz. The diagram below shows the method

My attenuator differed from G3WDG's in that used a 1, 2, 2, 5, 10, 20, 20dB sequence achieve 1dB steps; it also used only pi-secti attenuators, with resistors selected-on-test. must admit I like the idea of using T- or p sections according to which gives closest preferred resistor values.

The attenuator was only intended to be us for accurate measurement at 30MHz, but th technique provides a way of improving its hig frequency performance quite substantially.



Values

	R_1	R_2
1dB	12//12	910
2dB	12	430
5dB	68//56	180
10dB	120//220	100
20dB	510//510	62

15.5 Power Meters

15.5.1 A Simple VHF/UHF Power Meter - G4CNV

The detector or power meter shown below is a remarkably accurate way of measuring power in the range of 100 mW to 2 W, up to at least 500 MHz. The meter measures the peak rf voltage across the load, minus the forward voltage drop across the diode. This is divided by $\sqrt{2}$ to give the rms value, and the power calculated from this using the familiar equation:

$$P = \frac{V^2}{2R} \qquad (V = \text{peak value} = \sqrt{2} * \text{Rms value})$$

For germanium diodes, such as the OA47, the relationship will therefore be:

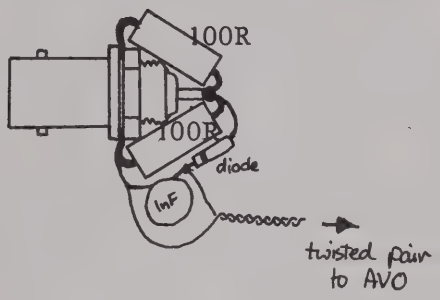
$$P = (V+0.25)^2/100$$

while for silicon diodes such as 1N914, 1N916, 1N4149, it is:

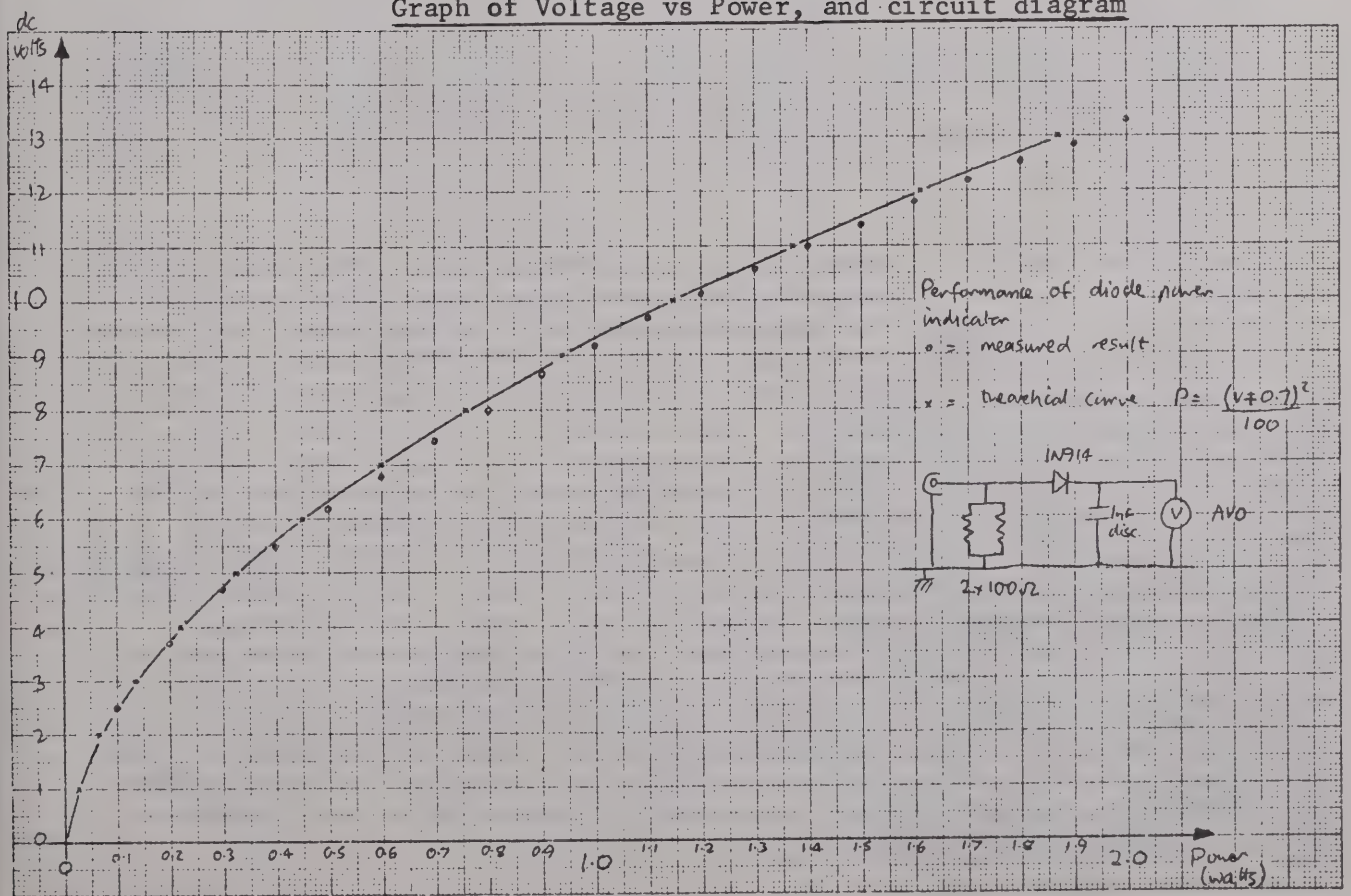
$$P = (V+0.7)^2/100.$$

The meter, originally suggested by G3VPF, could be used to measure the output from a 384 MHz oscillator board during alignment.

Construction using a BNC socket



Graph of Voltage vs Power, and circuit diagram



A Broadband Linear RF Power Meter

G3YGF

The circuit described here is not particularly accurate in absolute terms, but has a linear law over a very wide power range at lower power levels (10nW - 1mW). Under certain conditions, the DC current flowing when a diode detector operates into a low resistance load varies linearly with the applied RF power over a wide range. The dynamic range obtained depends greatly on the type of diode used, and is limited at the high power end by non linearity which will depend on the diode type, and at the lower end by thermal drift in the diode and DC amplifier used to measure the current. For the diode specified, the RF source impedance needs to be about 50ohms, and the DC load a zero impedance.

The diode required is a Zero Bias Schottky, Alpha types DDC4561 (1N23 type package) or 4562 (wire ended) which have a forward voltage of about 80mV at 1mA. The deviation from linearity is a maximum of about 1dB at the upper limit, and is considerably better at lower levels. Diodes with a higher forward voltage such as the HP HSCH3486 (160mV at 1mA) will not be so linear at high power levels (100uW-1mW) but this can be partly compensated for by using a resistor in series with the DC output to the op amp; For this diode, the resistor should be 180 ohms. Point contact diodes which have a forward voltage of 200-300mV at 1mA (eg OA90, 91 etc) can also be used, but they have a much lower limit of linearity at about 1uW, so will only give a linear dynamic range of 20-30dB. Other types with higher forward voltages are not suitable for use in this circuit.

Several types of microwave diode have been tested, such as the CS10B, and the Zero Bias Schottkys mentioned, and they seem to give a flat response (+-1dB) up to about 2GHz. I have not been able to check the sensitivity at higher frequencies, but it should remain fairly constant; the linearity certainly does.

Applications for this circuit include calibration of attenuators, measurement of received signal to noise levels, ground or sun noise; as a power indicator on a directional coupler power monitor, or connected to an aerial for measuring rf levels over short ranges. It is also capable of giving an output proportional to power in dB which would be particularly useful for plotting filter characteristics when combined with a tunable or swept oscillator or when needing to get a large dynamic range on the meter or scope display. It is hoped to produce modifications to the circuit shortly to make the meter read directly in dB.

Diode Mount:

The details of the diode mount are shown below. The upper frequency limit will be set by the method of construction, so obviously all components should have the shortest possible leads. The lower limit is set by the value of the smoothing capacitor, but if it is physically too large, it may degrade the high frequency performance. The components should be soldered directly to the body of the socket, or to the copper board next to the socket if it is mounted through some double sided PCB.

For best results a chip resistor should be used for the 50 ohm load, but below 1GHz there is little difference if a miniature film resistor with very short leads is used. Bear in mind that the centre pin of the socket can move slightly, so do not solder delicate components such as the chip resistor to it directly but use a short link of wire.

The SWR of this detector using a chip resistor is shown below, and is less than 1.2:1 at all power levels: it becomes slightly worse as the power is increased when the diode impedance decreases and begins to shunt the resistor. The return loss is only a couple of dB worse at 1GHz if an ordinary resistor is used. A diode in a standard waveguide mount should also be satisfactory, provided the resistance in the DC return

is less than about 50 ohms. The power-current characteristic is also shown below. A DC return in the signal source in not necessary for the circuit to work, but for best linearity the source shoud be around 50 ohms both at DC and the RF frequency.

Current Meter:

The circuit presents a very low input impedance to the diode and produces a voltage output either proportional to power, or log of power. In the linear mode, it gives an output of 0 to 1V for current ranges from 10nA to 10mA FSD, the sensitivity being selectable in decade ranges; in the log mode, the first op amp uses the exponential characteristics of the base - emitter junction of a silicon transistor to produce a voltage between 200 and 700mV, proportional to the log of the power, a 10dB change in power corresponding to a change of approximately 60mV. The exact value will depend on the transistors used.

The output can be zeroed using the offset null pot on the first op amp. This control should be on the front panel as it will be needed when working at low input powers. One position on the range switch shows the battery voltage, and has an FSD of 12V.

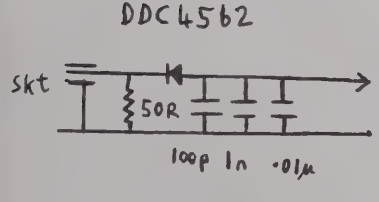
The second op amp is a unity gain inverter to enable a positive output to be obtained when using diodes mounted with either earthed anode or cathode.

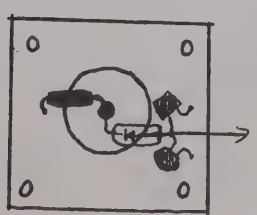
At low input levels in the log mode the meter reading may fluctuate wildly, or go negative; this is due to the input signal going very near to or through zero; this behaviour is normal, and can be avoided by adjusting the set zero control.

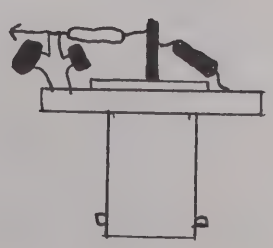
The values of the smoothing capacitors across the diode output or amplifier input may need to be reduced if a rapid response is required, as would be the case when sweeping a signal source, or increased if very low RF frequencies were being detected, eg <100kHz. There is no reason why it should not be used at AF. The low pass filter at the input cuts off at about 50kHz.

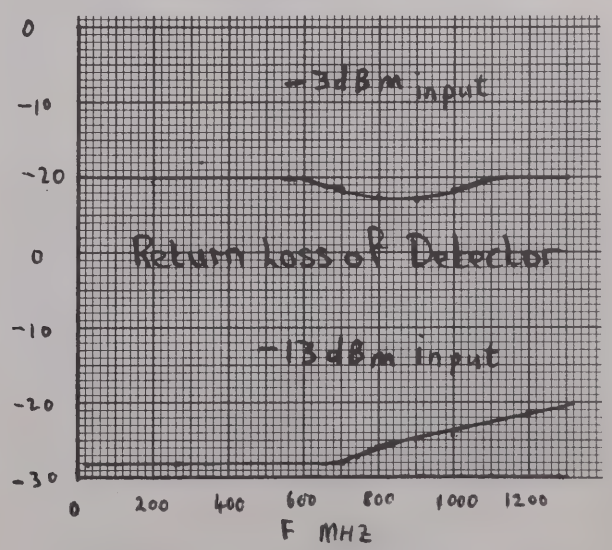
A FET op amp is used for IC1 for its low offset current, but the other op amp is not critical. It would be better to use FET ones throughout as their output can go very close to the supply rail, so the circuit will be less affected by low supply voltages. The Zero Bias Schottky diodes can be obtained from Mr R Evans, Alpha Industries, RMC House, Station Road, Witney, Oxon, at £7.35 for the DDC4561 (1N23 pack) and £8.66 for the DDC4562 (wire ended).

The detector and current meter are best mounted in the same box, and a battery supply used for the op amps. This helps avoid earthing problems which can occur as the input is very sensitive, being connected directly across the inputs of the op amp. A socket can be provided to connect the op amp input to an external detector, but coax cable must be used for this lead to prevent it from picking up and detecting stray RF.

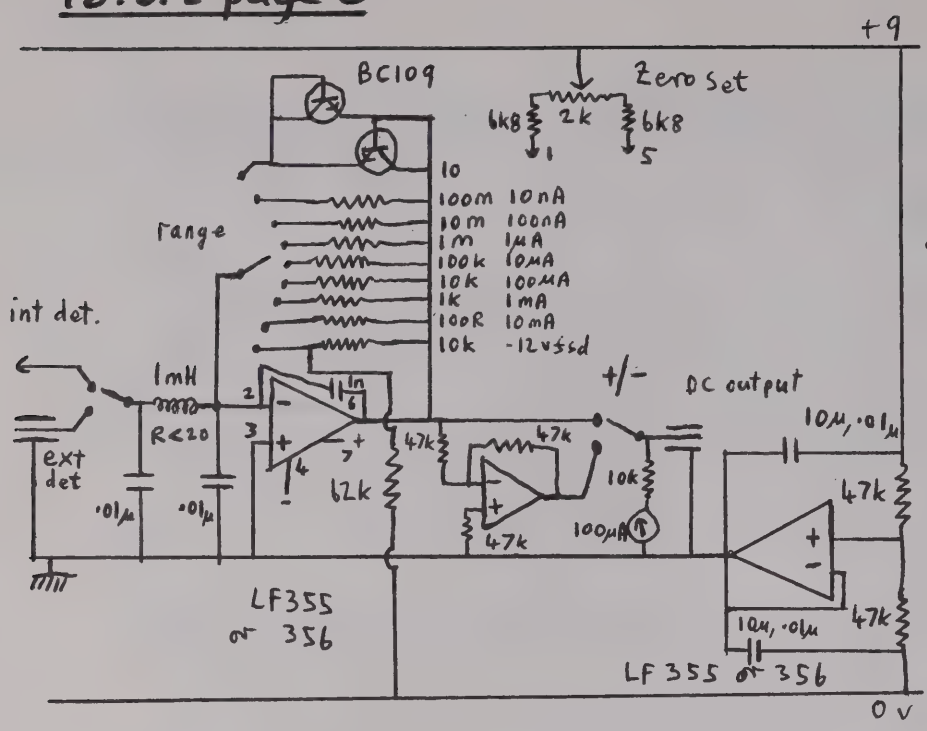


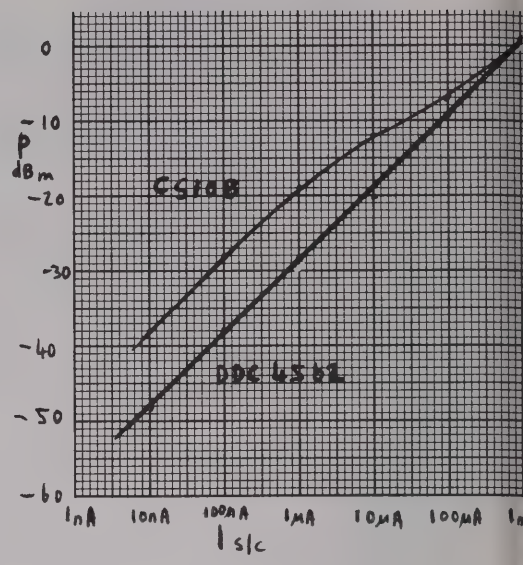






15.5.2 page 3



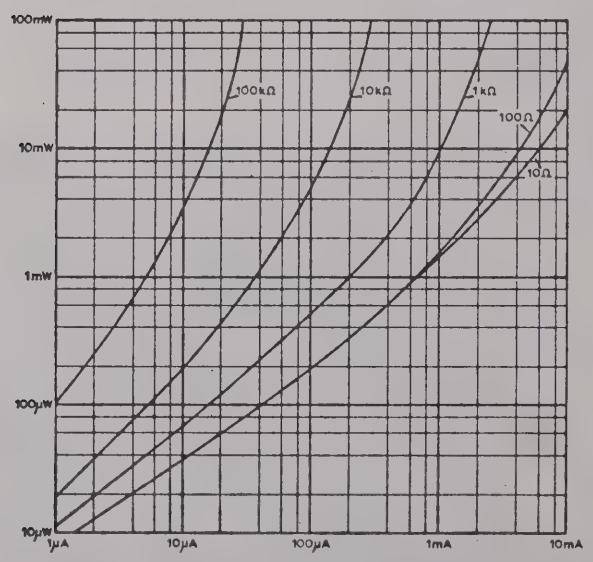


15.5.3

Rough measurements of power

Conventional point contact mixer diodes can be used to provide an indication of rf power with an accuracy sufficient to prevent gross mistakes in the construction of microwave equipment. The relationship between rf input and dc output for a typical diode is shown in the figure, and was derived from data given in the 1972 edition of the *Microwave Engineers Handbook*. Although this data applies specifically to X-band, the results obtained at other microwave frequencies would be expected to be similar. Each curve corresponds to a particular *total* series resistance. The resistance of meters, which is typically 100Ω for 1mA fsd, 1,000Ω for 100μA fsd and so on, in some cases is significant and must be allowed for.

As an example of the use of this data, suppose a mixer is fed via a 20dB (×100) directional coupler from a klystron, and a current of 0·9mA is observed on a meter having a resistance of 100Ω. The power indicated by the diode corresponds to 1·4mW of rf, which implies that the klystron is delivering in the region of 140mW. Comparing this value with manufacturer's data will give a guide to the efficiency of the system.



The approximate relationship between rf power input and dc output for a conventional point contact mixer diode

extracted from RadCom

15.5.4

Power/current relationship in Schottky diodes

The approximate relationship between the rf input power to a Schottky diode and the dc output is shown in Fig 2. The corresponding curve for point contact diodes is given in the September 1973 issue (p618). Note that each curve corresponds to a particular *total* series resistance. The resistance of a meter used, which is typically 100Ω for 1mA fsd, 1,000Ω for 100μA fsd and so on, may form a significant part of the total and must be allowed for.

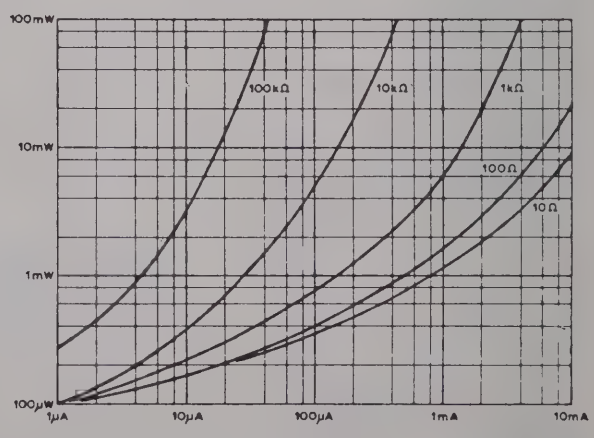


Fig 2. The approximate relationship between rf power input and dc output for a typical Schottky barrier mixer diode

extracted from RadCom

15.6.1 A Directional Coupler using Semi-rigid Coax - G3YGF

This design is an attempt to produce a coaxial directional coupler for use from HF to the low microwave bands. It's main feature is a fairly high directivity without relying too much on high precision construction.

The directivity obtained depends on the accuracy of the Zo of both the main and coupled lines, as well as the load terminating the coupled line. The directivity is obtained by sampling a small fraction of the E and H fields in the main line using a small slot which is less than a quarter wave long. These fields leak out into the coupled line and reproduce the forward and reflected waves in the coupled line. The accuracy of the sampling is improved in this design by using equal diameter lines with the same dielectric.

Consider Fig 1, where the coupler has a load with a 1.2:1 swr (20dB return loss) on it. The forward power will couple -30dB into the termination, and the reflected power -50dB into the detector. If the coupled line termination has an swr of 1.2:1, then this will also reflect -20dB of the incident signal (-30dB), ie -50dB back to the detector. This is then producing a signal at the same level as the one to be measured.

One of the main problems in obtaining a high directivity is that of making the lines, either coax or stripline sufficiently accurately, and also doing the transitions from them to the loads and connectors. For this reason, any couplers built using homebuilt lines are unlikely to be very directive unless they are checked or adjusted on known testgear.

Construction

Semi rigid coax of any diameter can be used, 0.141, 0.250 and 0.325" being common values. They can be used with bnc, sma or N type connectors according to the size, but ensure that the connector is intended for use with that size of cable, or there may be a significant swr introduced. The larger sizes are stronger mechanically, and easier to work with, and can be made to fit into N connectors quite easily.

Hold the cable in a vice and file away the outer to form a flat on the cable of the desired length. It should be a quarter wavelength long in the cable dielectric (ptfe) at the highest frequency required. (e_r is 2.1) Fig 2.

Be careful not to tear the copper outer away from the ptfe by over enthusiastic filing. At each end of the slot taper the thickness of the copper so that the gap between the two cables is as small as possible. Bend the second line to the shape shown before filing the slot in it. Then clamp the two together and solder them, all the way round the join to seal the coaxes again. The slot width should be about 1mm on the 0.141, and 2mm on the larger sizes.

The high directivity is only obtained for fairly weak coupling, and the slot widths shown give about 35dB maximum at the high frequency end of their range. 30 dB should be regarded as the maximum value to be used, as the slots then become too wide and begin to alter the Z_o of the line.

The coupling coefficient and directivity vary with frequency as shown in Fig 3 for a typical example. A directivity of 20-25dB is typical, and the coupling varies at the rate of 20dB per decade of frequency.

Connectors will be required at the ends of the main line, but on the coupled line there are two options: either use connectors, which is the most flexable option as then its performance can be measured easily on test equipment. Alternatively, the load and detector can be soldered directly onto the cable. A 50 ohm chip resistor soldered across from inner to outer will provide a good enough load. The diode detector can then be soldered across another 50 ohm chip resistor at the other end of the line.

The type of detector must be chosen according to the power level and coupling coefficient used, as shown in Fig 4. A receiver or amplifier is needed at very low power levels, the diode short circuit current meter previously described in the newsletter is useful in the range -40 - 0dBm, and an open circuit voltage detector (described in the newsletter and the Microwave committee 384MHz oscillator article in Radcom) is suitable for powers above 0 - 10dBm. Above about 20dBm the dissipation in the chip resistor becomes excessive. Here, external attenuators could be used. The coupler would normally be used as shown, but provided that the swr of the resistor and detector are good enough, they can be mounted at both ends of the coupled line so that power in both directions can be monitored simultaneously

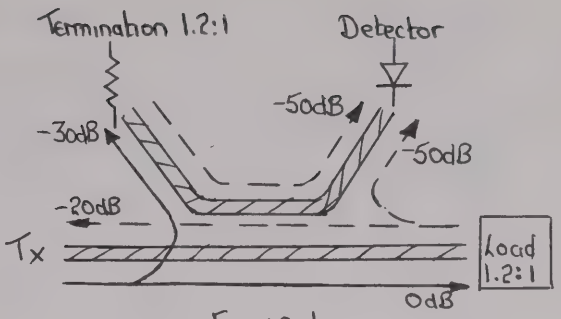


Figure 1

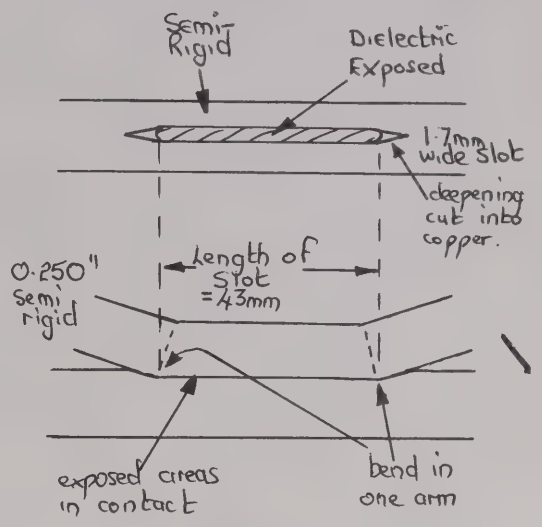


Figure 2

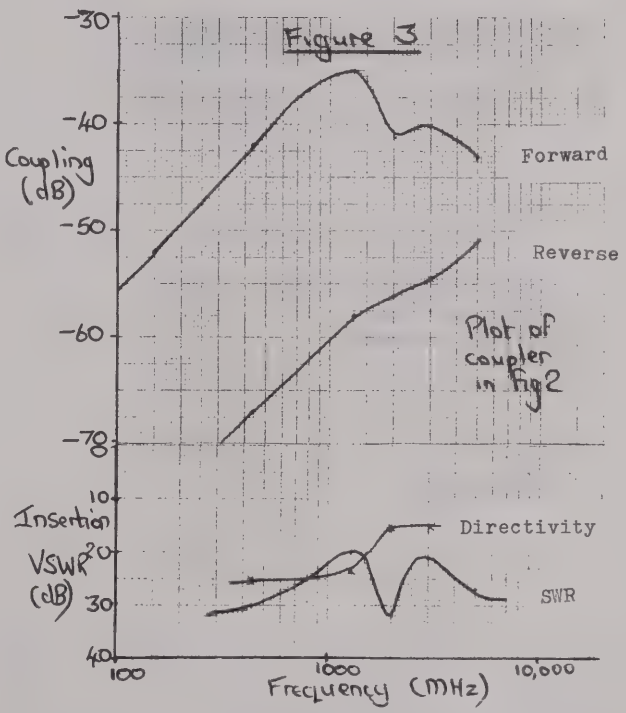


Figure 3

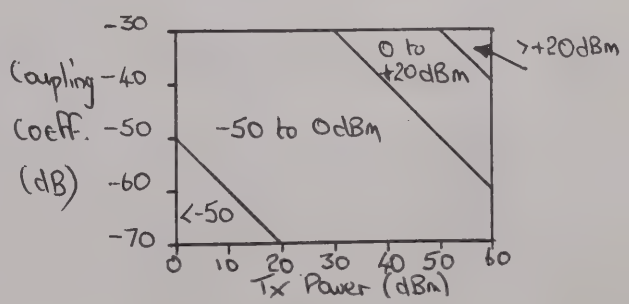


Figure 4

15.6.2

High Power Directional Coupler for 1296MHz

This design is taken from the Swedish West Coast Microwave Group newsletter and avoids tying up a BIRD or similar wattmeter for monitoring the power output from an amplifier. It was originally intended for use on 70cm with an amplifier such as the K2RIW but will work quite well on 23cm and would be useful to monitor the output from a 2C39 PA, for example.

The components required are: a length of 0.141" semi-rigid coax, a 47 ohm resistor, two pieces of copper or brass tubing and coax connectors. The two lengths of tubing are chosen with diameters to produce a 50 ohm line (see diagram) and suitable values are 19.5mm inner diameter for one length and 8.5mm outer diameter for the other. The appropriate equation is $Z=138\log D/d$.

The slot in the outer length of tube can be made by drilling a few holes in a line and then joining them up with a file to produce a short length of slot. The slot is then extended with a small hacksaw blade to the full length. Obviously this operation could be simplified if a milling machine is available

The semi-rigid coax is then filed flat for a length equal to the slot length. The outer jacket if filed away to expose the teflon dielectric to a width of about 2.5mm. This can now be soldered in place on the outer length of tube so that the two slots are aligned.

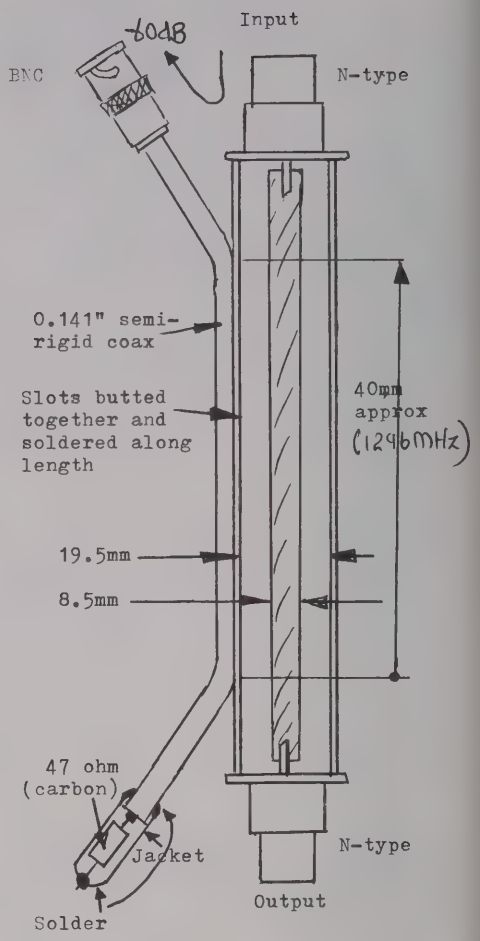
Prepare both ends of the semi-rigid as for BNC connectors. On one end a BNC (male or female) is fitted in the normal manner. On the other end a 47 ohm load is connected between the inner and outer. Firstly the resistor is soldered to the inner with as short a lead as possible. Then a piece of copper jacket is slipped round the resistor and soldered against the end of the outer of the semi rigid (see diagram). Finally the other end of the resistor is soldered to the jacket, the jacket being compressed first so that the end can be completely sealed.

All that now remains is to fit the inner length of tubing and the connectors in place, care being taken to install the tube in the centre.

The indicator for the coupler can be an RF power meter or simply a diode detector with a meter. Calibration can be done with a receiver, signal source and a calibrated attenuator. Alternatively a direct comparison with a calibrated BIRD or other power meter can be used.

The coupling is about -60dB for the size of slot given on 1.3GHz. Making the slot larger will increase the coupling, as will a smaller diameter mainline. The directivity was measured at 18dB on 1.3GHz. An alternative would be to build the coupler as a dual coupler so that both forward and reflected power can be monitored. For use on 70cm a longer length of slot is recommended (to give a sufficiently high coupling) of about 3x the 23cm length. The exact length is, however, uncritical.

Thanks to the writers, SM6AOM, SM6ECR and to SM6CKU for the information.



15.6.3

Notes on using Directional Couplers - G3WDG

G3WDG has calculated error limits for a given directivity of coupler when using it to measure the return loss (or VSWR) of an unknown load. These are summarised in the table below:

Measured Return Loss (dB)	Range of True Return Loss (dB)	
	20dB Dir'y	25dB Dir'y
10	7.6 to 13.3	8.6 to 11.7
12	9.1 to 16.4	10.2 to 14.2
14	10.5 to 20.0	11.8 to 16.9
16	11.8 to 24.7	13.4 to 19.8
18	12.9 to 31.7	15.0 to 23.1
20	14.0 to ∞	16.1 to 27.2
22	15.0 to 33.7	17.3 to 32.7
24	15.8 to 28.7	18.5 to 43.3
26	16.5 to 26.0	19.5 to 44.3
28	17.1 to 24.4	20.4 to 35.7
30	17.6 to 23.3	21.1 to 32.2
∞	20	25

For example, if a coupler with 20dB of directivit is used to measure the return loss of an antenna and the measured value is 26dB then the true return loss can be anywhere between 16.5 and 26.0dB. Furthermore, attempting to set up the antenna for zero reflected power will simply result in an antenna whose return loss is 20dB (or VSWR 1.2 to 1). This is what is usua what is done in practice - the antenna adjusted for zero reflected power and it accepted that its return loss is the same as the directivity of the coupler. In doing this it may be necessary to insert different lengths of cable between the reflectometer and the antenna under test.

An alternative approach is to adjust the return loss to that of the directivity of the coupler such that inserting lengths of cable between the antenna a coupler does not change the indicated reflected power Under these conditions the antenna would then be perfectly matched. However this method is much more laborious and assumes that the directivity of the coupler is accurately known.

Section 16 Miscellaneous

EME Link Budget Program - G3WDG

The following BASIC program was supplied by G3WDG and calculates the signal to noise ratio for an EME link. All the REM statements may be ommitted if desired. The noise figure includes sky noise and ground noise as well as receiver noise. Two different antennas may be given for a test between two stations, or for echo testing enter the same antenna twice.

```
100 REM THIS PROGRAM CALCULATES THE SIGNAL TO NOISE
110 REM RATIO OF EME SIGNALS, GIVEN THE EQUIPMENT
120 REM PARAMETERS, FOR THE AVERAGE EARTH MOON
130 REM DISTANCE. AT PERIGEE SIGNALS WILL BE 1DB
140 REM STRONGER, AT APOGEE 1DB WEAKER.
150 REM
160 REM PROGRAM INPUTS ARE OBVIOUS: NOTE THAT THE
170 REM RECEIVER SENSITIVITY MAY BE GIVEN AS EITHER
180 REM NOISE FIGURE [NUMBER < 20] OR NOISE TEMP
190 REM [NUMBER > 20]; ANTENNA PARAMETERS ARE EITHER
200 REM DISH DIA IN FEET [ENTERED AS A NUMBER], OR
210 REM ANTENNA GAIN [ENTERED AS GAIN/100, E.G 9DB
220 REM ENTERED AS .09].
230 REM
240 REM WRITTEN BY C.W. SUCKLING G3WDG, 28 FEB 1983.
250 REM
260 REM GET INPUT DATA
270 GOSUB 350
280 GOSUB 370
290 GOSUB 390
300 GOSUB 410
310 GOSUB 430
320 GOSUB 450
330 GO TO 480
340 REM SUBROUTINES TO GET INPUT DATA
350 PRINT "SYSTEM NOISE TEMP OR NOISE FIGURE "; \ INPUT N
360 RETURN
370 PRINT "RX BANDWIDTH (HZ)              "; \ INPUT B
380 RETURN
390 PRINT "FREQUENCY (MHZ)                "; \ INPUT F
400 RETURN
410 PRINT "ANT 1 DIA (FT) OR GAIN (.DB)   "; \ INPUT D1
420 RETURN
430 PRINT "ANT 2 DIA (FT) OR GAIN (.DB)   "; \ INPUT D2
440 RETURN
450 PRINT "TX POWER (W)                   "; \ INPUT P
460 RETURN
470 REM ANTENNA CALCULATIONS
480 D=D1 \ GOSUB 530
490 G1=G \ A1$=D$ \ A1=A
500 D=D2 \ GOSUB 530
510 G2=G \ A2$=D$ \ A2=A
520 GO TO 590
530 IF D<1 THEN  GO TO 560
540 D$="FT" \ G=(4.3422*LOG(3.094*D*D*F*F))-60 \ A
550 GO TO 570
560 D$="DB" \ G=100*D \ A=G
570 RETURN
580 REM CONVERT NOISE FIGURE TO NOISE TEMPERATURE
590 IF N<20 THEN  GO TO 620
600 T=N \ N$="K"
610 GO TO 640
620 T=290*(EXP(.2303*N)-1) \ N$="DB"
630 REM CALCULATE RECEIVER NOISE POWER
640 R=-228.6+4.3422*LOG(T)+4.3422*LOG(B)
650 REM CALCULATION OF PATH LOSS
660 L=-271+8.6844*LOG(1296/F)
670 REM CALCULATION OF SIGNAL TO NOISE RATIO
680 S=G1+G2+4.3422*LOG(P)+L-R
690 S=INT(S*10)/10
700 REM PRINT OUT RESULTS
710 PRINT
720 PRINT "CODE  PARAMETER          CURRENT VALUE"
730 PRINT
740 PRINT "1     NOISE TEMP/FIG   ";N;N$
750 PRINT "2     BANDWIDTH        ";B;"HZ"
760 PRINT "3     FREQUENCY        ";F;"MHZ"
770 PRINT "4     ANTENNA 1        ";A1;A1$
780 PRINT "5     ANTENNA 2        ";A2;A2$
790 PRINT "6     TX POWER         ";P;"W"
800 PRINT
810 PRINT "SIGNAL TO NOISE RATIO ";S;"DB"
820 PRINT
830 REM GET FURTHER INSTRUCTIONS
840 PRINT "NEW PARAM(1 TO 6), RE-RUN(R) OR QUIT(Q)
850 INPUT A$
860 PRINT
870 IF A$="Q" THEN STOP
880 IF A$="R" THEN  GO TO 270
890 IF A$="1" THEN  GOSUB 350
900 IF A$="2" THEN  GOSUB 370
910 IF A$="3" THEN  GOSUB 390
920 IF A$="4" THEN  GOSUB 410
930 IF A$="5" THEN  GOSUB 430
940 IF A$="6" THEN  GOSUB 450
950 GO TO 480
960 END
```

Functions used: LOG - natural log, EXP - exponent and INT - truncate to an integer

Here is the result of a program run using a 13ft dia dish on 1.3GHz:

CODE	PARAMETER	CURRENT VALUE
1	NOISE TEMP/FIG	1.3 DB
2	BANDWIDTH	100 HZ
3	FREQUENCY	1296 MHZ
4	ANTENNA 1	13 FT
5	ANTENNA 2	13 FT
6	TX POWER	350 W
	SIGNAL TO NOISE RATIO	6.2 DB

A few multiple statement lines have been used - the character "\" used to separate statements. On many machines these will need to be changed to ":", or the statements may be put on separate lines. Also, on some machines PRINT and INPUT statements may be combined into one statement. Many tnx to Charlie for the info.

16·2

Heliax connectors using standard N-type plugs and sockets

More and more operators are realizing the advantages to be gained in using very-low-loss coaxial cables on the lower microwave bands. Recently, reasonably large quantities of Andrews FHJ-4 cable (a 0·5in diameter foam-filled heliax cable) have become available. This is very useful indeed, having losses of only 2dB/100ft at 432MHz, 4dB at 1·3GHz and 5·5dB at 2·3GHz. The writer has found it to be quite useful even at 10GHz, where the losses are less than 2dB/10ft.

The main difficulty in using this cable up to now has been the supply of suitable connectors, which are relatively expensive when new. G3YGF has supplied details of a method for adapting standard N-type connectors to fit this cable. The resulting fittings have been tested for loss and swr up to 10GHz, and have been found to perform as well as the proper connectors.

There are two common types of N-type plug, and only the shorter, thicker version is suitable for modification. To check a particular connector, measure the overall length of the plug (with the cable clamping nut removed) and the diameter of the plug body (not the knurled nut). Suitable connectors are 1·32in long and 0·75in in diameter, while those which cannot be modified are 1·52in in length and have a diameter of 0·685in. The modifications should also work for cable mounting sockets, although G3YGF has not yet encountered a socket of the correct type. The parts of the plug which are used are the main body, cable clamping nut, centre pin, and the two ptfe washers which locate the centre pin.

The first stage is to bore out the cable clamping nut to 35/64in id, preferably on a lathe. The nut can be screwed into the body of the plug during this operation, for support. The second stage is to manufacture an adaptor piece to allow the centre pin to be fitted on to the centre conductor of the cable. This adaptor is shown in Fig 1. It is recommended to construct this out of copper, rather than brass, as this allows some flexibility during alignment, with less risk of the centre pin breaking off. Start by turning down a piece of copper rod to 0·28in diameter, and turn the end pin to size. The 45° taper at the end of the pin is then cut, using a chisel shaped tool, at 45°. Next, part or cut off the piece 0·2in from the large end of the taper, and refit in the lathe the other way round. Face off the cut end to the 0·19in dimension. Cut a 45° taper on this end, in the same way as before. Bore out the 0·175in diameter hole in this end, taking care not to go deeper than shown, or the walls near the taper may get excessively thin and thus weaken the adaptor.

The rear ptfe washer should be filed slightly to reduce its diameter, and its front edge chamfered, so that it is an easy fit inside the body of the plug and will not jam at an angle inside the plug during assembly.

The end of the cable should then be prepared as follows. Remove the plastic sheath 1in from the end. Mark the outer conductor at a point 0·15in from the end, and cut through all the way round, going into the foam a short distance. Remove the outer and foam by twisting. Clean the centre conductor, being careful not to scrape through the thin layer of copper.

The complete connector may then be assembled, referring to Fig 2. Tin the inside of the cable clamping nut, the pin and the inside hole of the adaptor. Mount the cable horizontally in a vice, tin the centre conductor, heat up the adaptor and solder to the centre conductor, being careful to push the adaptor on as far as it will go. Assemble the rear ptfe washer on the connector pin, and solder the pin to the adaptor. The larger ptfe washer may then be placed in position on the pin.

The clamping nut is then positioned over the outer conductor, and the plug body screwed on to the nut fairly tightly by hand. Ensure that the pin is in the centre of the plug, and adjust if necessary by bending carefully. Slide the plug along the cable so that the centre pin is recessed in the plug by the correct amount. Hold the nut firmly in position on the cable, and unscrew the plug body. Without moving the nut, solder it on to the outer. Then reassemble the plug and check that the centre pin is still in the correct position, with the plug screwed up tightly. If it is not quite correct, some adjustment is possible by resoldering the centre pin on the adaptor in a different position.

Before reassembling finally, remove all traces of solder from the foam dielectric and cut out the dielectric to a depth of about 0·1in.

Most of the soldering can be done with a 50W iron, but the nut on the outer may need a small gas torch. This operation should be done fairly quickly to avoid melting the dielectric excessively. The finished plug may be waterproofed with self-amalgamating tape, mainly to prevent water entering the plug through the threaded portion.

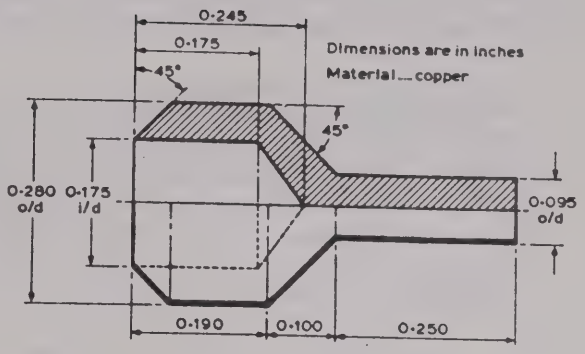


Fig 1. Centre pin adaptor

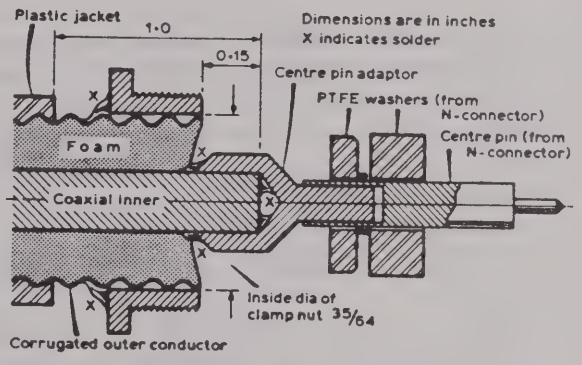


Fig 2. Assembly of plug

extracted from RadCom

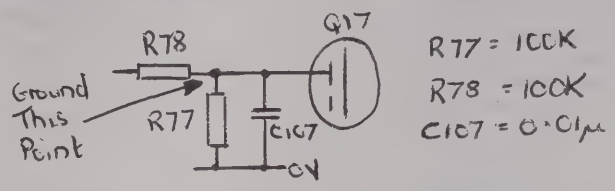
16.3 Notes on the IC202

IC202 Spurii on 1.3 GHz

Those of us who use an IC202 as a tunable IF for 1296 may have noticed a spurious signal in the beacon part of the band. This occurs at about 1296.86 MHz, tunes very fast, and is most noticeable then the IC202 is near the 1.3 GHz antenna.

It is actually caused by the 87'th harmonic of the IC202 vfo fundamental, which comes out at 1296.86 MHz, and is radiated from the IC202 at a suprisingly high level (a test dipole held near the rig can easily pick up -70 dBm). The fundamental is at 14.907 MHz, and this is tripled twice to provide the IC202 LO, and it is a harmonic of the first multiplication (29*3*14.9) that causes the problem. Has anyone noticed the same effect on other 144 MHz rigs with a 10.7 MHz IF?

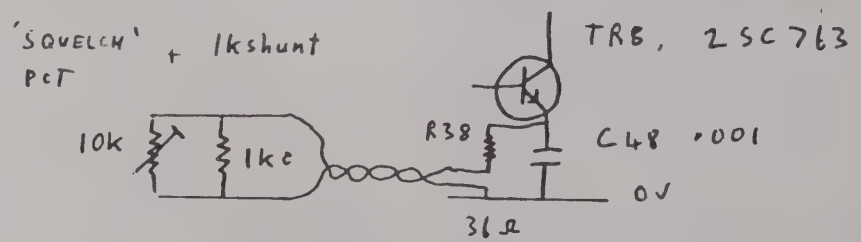
Here is an idea from F6DLA to modify an IC202S to reduce the power output when driving 10GHz transverters. It consists of shorting the Gate 2 of the 1st 144MHz transmit amplifier to ground, see Fig, This reduces the output power to about 100mW which reduces the battery drain, and wastes less power in the attenuator. This is done by a small jumper lead, which fits into two small sockets, one soldered to the screen round this part of the circuit, and the other, to one end of the 100k resistors by scraping some of the white insulation off the lead. This avoids having to remove the pcl



Petra, G4KGC reports that the tuning range of an IC202 can be increased considerably merely by altering the trimmers provided for the existing crystals. A range of about 1MHz can be obtained with one crystal, although the scale is then very non linear, each 10KHz division then varying from about 30kHz to 70kHz at each end of the scale. This mod is useful to overcome the limitations of a restricted tuning range on a 10GHz IF, for example.

Power reduction for 2m Transceivers

G3RZD has sent in details of how he reduces the power output from a liner 2 transceiver to drive the JVL mixer. The emitter resistor R38 (36R) is unsoldered from board and taken to earth by a twisted pair and variable resistor as shown in the diagram.



IC2Ø2's

G4CNV (and others) have tried the front-end modifications suggested in DUBUS-INFO 3-80. The first of these, which consists of bypassing a stage of front-end filtering, is trivial to carry out, and improves the sensitivity by about 3dB. The second modification, which reduces the insertion loss of a filter between rf stage and mixer, is more difficult to implement, and gives much less improvement, so probably isn't worthwhile. Obviously, reducing front-end selectivity has its drawbacks, but in general the strongest off-channel signals to be encountered will be in-band ones, where front end selectivity is no help.

16.3 page 2

Odd Behaviour of the IC202 G3YGF

A popular choice of crystals for the IC202 consists of 144.0 - 144.2, 144.2 - 144.4, 144.4 - 144.6 and 144.8 - 145.0. This combination of the last two crystals is not recommended by Icom, but the reason for it is not widely known, and it may appear to perform ok.

The problem is that the top two crystals interact at one point in the .8 - 145 .0 vxo range. The frequency jumps by about 10 or 20 kHz as the frequency of approximately 144.850 is approached from either direction. Thus there is a small range of frequencies which it is not possible to reach! The fault is rather subtle, as there is no audible sign of the jump unless there is a signal in the passband at the point where it jumps. It was noticed when the rig was used as an IF for 1296MHz while listening for beacons - one around that frequency proved very elusive, as it only appeared when the drift in the converter crystal brought it outside the 'forbidden band'.

The only cure I have found is to remove the .4 - .6 crystal when listening around this frequency. The effect can be found by slowly tuning a signal generator across the .8 - 145.0 band, keeping it in the passband with the 202 tuning, and listening for the jump. Alternatively, put the rig on transmit and observe the frequency on a counter.

16.4 G3JVL has discovered some (unknown?) logic in the WR/UR numbering system for waveguides. The number is the inside dimension of the broad face of the guide in 0.010 inches. ie UR90 is waveguide 16, whose inside dimension is 0.9 inches... obvious once you see it! It also tells you the cutoff wavlength, as this dimension is a half of the free space wavelength at cutoff.

16.5

Problems in using Droitwich-locked Frequency Standards -G3YGF

The following notes relating to the 200kHz Droitwich transmission were obtained from the BBC Research Department, Kingswood Warren.

1. The programme modulation rolls off below 40Hz.
2. In the 25Hz bandwidth centred on the carrier, the residual am and psk due to programme material are 2% and 2° respectively.
3. There is am-to-pm conversion in the transmitter, approx. 17° of phase shift for 100% modulation.
4. There is now some slow psk data on the carrier, approx. ±22° shift at a 25Hz rate, using some sort of random code.
5. There is a slave transmitter near Lossiemouth (Dundee) which locks off-air to the main Droitwich signal. It makes corrections to itself in the form of 30nS jumps in the phase of the carrier, approx. every 8 seconds.

Using the Droitwich signal as a reference for a phase-locked vhf oscillator which is multiplied to 10GHz as a frequency marker, these effects show themselves as follows:

a) AM-to-pm conversion in the tx causes the frequency to pull approx. ±1/2kHz on loud speech/music peaks.

b) The psk data spreads the carrier out over several kHz when it is present.

c) The slave transmitter jumps cause the 10GHz signal frequency to jump several kHz, then slowly return to the mean value. This may happen several times a minute. In S. England this is most noticeable after the main transmitter has closed down at night. It may be a more serious problem during the day for those served by the slave transmitter.

d) Lastly (and probably least!), am-to-pm conversion in your own 200kHz rx limiting i.f. strip can cause problems similar to a). The relative merits of limiting and crystal filtering for a 200kHz rx are still debated, but in view of the inherent am-to-pm conversion of the transmission, limiting has a lot in its favour, especially since you can use log(signal strength) to give an audio output to listen to the transmission as well!

16.6 UoSAT.

Control of the satellite was regained in late Sept and it was hoped to stabilise it by December. There are two microwave beacons on board: 2401.00MHz and 10.426GHz beacons. The 2401MHz beacon is running 125mW to a LH circular helix, 6.5dBi gain and is modulated at 10kHz deviation. The path loss when the satellite is overhead is 152dB, and when on the horizon is 169dB. The 10GHz beacon is also running 125mW, LH circular polarisation from a 4 turn slot helix which has 8dBi gain. Path losses on this band are 169dB and 181dB. No details (as far as we know) are available on when they will be qrv.

Notes on listening for the UoSAT Beacons

The 2.4GHz beacon is crystal controlled and may have up to ±50kHz doppler shift on it. Calculations suggest that it should be audible on an omni on most orbits. For example, using a 3dB NF, 1kHz bandwidth, 0dBi ant, the S/N overhead will be 18dB, and on the horizon 1dB.

The 10.4GHz beacon is a free running oscillator using a dielectric resonator, which is probably similar to a Gunn in bandwidth. An IF bandwidth of 5 to 20kHz is probably appropriate. The doppler shift may be up to ±220kHz. The satellite will pass through the beamwidth of a 2ft dish in about 10 seconds in the worst case, so unless a good tracking system is built, a lower gain antenna eg a horn seems more likely to produce success. For example, Using a 23dBi horn, 9dB NF, 10kHz bandwidth, the signals will range from 10 to -2dBN. It should not be too hard to track manually with a 23dBi horn.

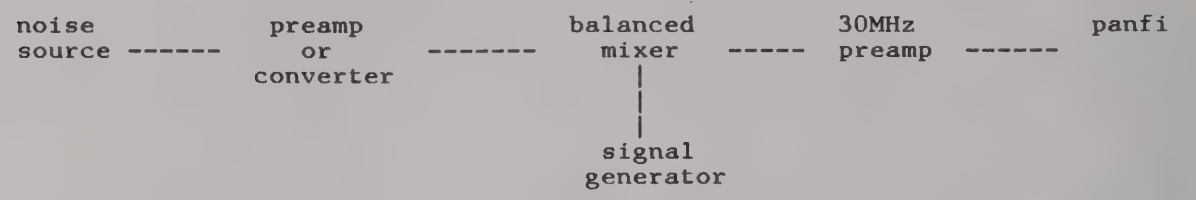
Both these calculations assume that the correct circularity polarised antenna is used. If a linear one is used, signals will be 3dB worse, and may fade even lower at times. It would be possible to use an array of 2.3GHz antennas to derive pointing information for the 10GHz beam.

Some pitfalls in noise figure measurements

(from experiences at the Midlands and Martlesham Conventions)

by G3YGF and G4FZZ, Oct 81

Instruments similar to the AIL Precision Automatic Noise Figure Indicator (Panfi), are commonly used to measure noise figures of preamplifiers and converters in conjunction with a broadband calibrated noise source as shown below. Many converters do not come out at 30MHz, so there is often more than one converter in the chain.



Unreliable, misleading, or wrong results can be obtained for several reasons:

a) The source impedance (SWR) of the noise source may change between its noise on and noise off conditions, thus altering the noise output from the amplifier regardless of the noise coming from the source.

This can be avoided by using an attenuator between the noise head and the device under test; a value of 10 - 15dB should remove the problem. The NF reading obtained must then be corrected by subtracting from it the value of the attenuator. This attenuation must be known quite accurately, and it must present a good match. Different results obtained with and without the attenuator indicate that the noise source SWR is changing; the value obtained using the attenuator should be the more accurate one.

b) There may be a large tolerance on the spec of the ENR of the noise source. (typ .1-.5dB) Nothing can be done about this uncertainty which limits the overall accuracy of the measurement. If an attenuator is used with the noise source, then the uncertainty in its value must be added to that of the ENR.

c) The contribution to the noise figure of the second and subsequent stages must be allowed for. This requires a knowledge of the noise figure of the system at the second stage, and the gain of the first stage. The noise figure of the system at the second stage should be fairly low, perhaps 3-5dB, and the input match must be reasonable to prevent it from encouraging instability in the items under test. Low value attenuators may be necessary between some devices to keep them tame.

d) Any of the preamps or converters may be unstable or oscillating at some frequency quite unrelated to the band for which they were designed. GaAsFETs or even transistors in VHF/UHF circuits are capable of hooting at many GHz. In the absence of a 'DC - light' spectrum analyser or detector, hysteresis or fluctuations in the device DC conditions or noise figure when bits of the circuit are prodded are good indications of trouble.

e) The overall gain may be high enough for the noise to drive the later stages of the chain into non linearity, or saturation. Remember that you may be working in bandwidths of 10MHz or more, so the broadband noise power mounts up quite quickly. Since small signal amplifiers may begin overloading at an output of -10dBm, ie an input of - 20 to - 30dBm, and as the thermal noise level in 1MHz bandwidth is - 110dBm, this leaves a relatively small range of gains over which the system will work reliably. The gain should just be sufficient to get the noise level about 20dB over the Panfi AGC threshold of -70dBm, ie about 60dB from preamp input to Panfi input. If the gain is excessive, reduce it by removing amplifiers rather than inserting attenuators. It is no good just cascading a selection of preamps and converters and hoping.

f) Any strong local transmissions on any of the IFs used or on the spurious responses may be picked up and will alter the noise ratios seen by the panfi. The panfi has a bandwidth of 5-10MHz, and most converters are in the range 1-10MHz wide. Thus interference is possible from quite a wide range of frequencies, including the local oscillators in other converters waiting to be tested! A spectrum analyser is really required to check that the IF output is clear of other signals. (This task would be a lot easier if a narrow IF such as an SSB receiver were used, as the frequency used could be monitored aurally).

The system is particularly sensitive to transmissions in the band being tested, as signal levels at the preamp input are around the thermal noise level, and any stray pickup must be about 20dB below this to avoid trouble. A surprising number of local signals can be heard on most receivers even with no aerial connected. Comparison of two measurements at slightly different frequencies within the system bandwidth may reveal unexpected differences.

g) In the absence of filtering, the balanced mixer may also perform mixing with harmonics of its local oscillator. Matters are made worse by the harmonic relationship between many of the amateur bands and IFs used.

The performance of such mixers also degrades in many respects if all its ports are not terminated in 50 ohms at all frequencies present; such correct termination is rarely the case.

Indeed, some converters have quite curious spurious responses; some 432 to 144MHz converters with a 288MHz local oscillator have a response which is only 13dB down on the wanted one, and is caused by the second harmonic of the local oscillator also mixing with the signal! eg 432 - 288 = 144; 576 - 432 = 144.

h) The balanced mixer, and some converters tested, well respond equally well on the image as on the signal frequency. The noise source is broadband, and depending on the position of any filtering, it is possible that the measured noise figure may be anywhere between the SSB and DSB values. It is essential to define the frequencies being measured by using effective bandpass filtering throughout the system.

i) Most converters leak significant amounts of local oscillator power (-20dBm), from their input and output sockets. This leakage from the output into a broadband IF preamp can cause overload effects in the preamp, reducing its gain for the noise signals. If either input or output are connected to a broadband mixer, then this signal is competing with the real local oscillator, and all manner of spurious products and effects are possible.

The main cause of trouble is that the levels and frequencies present are rather uncertain when a test bench is set up for a special occasion with unfamiliar components in an unknown RF environment, and the stability, gain or frequency response of the item being tested are also unknown variables.

With some additional testgear these problems can be avoided.

1) A spectrum analyser can be used to examine the signal levels and frequencies present at various stages through the chain, and to detect the presence of instability or RF breakthrough. At the very least, some form of broadband power detector should be available.

2) Gains or frequency responses can be measured on a network analyser, or using a signal generator and power meter.

3) Efficient filtering is vital to avoid overloading and spurious response problems; a decent band pass filter is needed between any mixer and the following preamp, and also one to attenuate image responses on the mixer input.

4) Unsteady readings on the Panfi indicate RF breakthrough or instability. They may correlate with movements of the people/equipment altering the RF pickup. Thorough screening of the equipment, or shutdown of the offending transmitters are the only cure.

5) Amplifier overload is indicated by a large increase in noise figure when, for example, a preamp is put in front of a converter, or when another preamp is put in front of the existing system.

6) Spurious signals in the passband can be indicated by the noise figure varying rapidly with frequency as the local oscillator or Panfi IF frequency is varied over a relatively small range; high order local oscillator products may move right across the passband for only a few hundred kHz change at the fundamental. When everything is working correctly, the noise figure should not vary greatly over the width of the IF - several MHz.

The use of purpose built converters for the bands required is strongly recommended, and avoids many of the problems of broadband mixers. Even so, there should be good selectivity before and after the mixer in every converter; this may either be already built into the converter, or have to be added as a separate filter.

Typical system for testing 1296MHz preamplifiers

noise source	1296 preamp	1296 bpf	conv	144 bpf	conv	28 bpf	preamp	panfi
gain	15dB	-3dB	+20dB	-3dB	+20dB	-3dB	20dB	
-110	-95	-98	-68	-71	-51	-54	-34	

(dBm in 1MHz)

The panfi requires a minimum input of -70dBm to operate its agc, and has a maximum input of about 0dBm. So the level 'noise off' should be 10-20dB above this agc level, but well below the maximum input level, as the noise source will increase the signal by up to 15dB.

Confidence in the measurements can only be obtained when the above checks have been performed, and the system is giving repeatable and consistent results with a variety of preamps. Be very suspicious of, and investigate, any unusual or unexpected behaviour.

Note: The text is based upon an article published in VHF Communications, 1/82.

16.8.1 page1

VHF Communications

The following list contains items of interest to microwave enthusiasts which were published in VHF Commmunciations during 1979, 1980, 1981 and 1982.

1/79 "A Transistorized Linear Amplifier for the 23cm Band" by DF8QK and DC0DA.
1/79 "A Transmit Mixer and Linear Amplifier for the 13cm Band Equipped with a 2C39 Tube" by DF5QZ.
1/79 "The 10GHz Amateur Band - Considerations of Present and Future Technologies - Part 2" by DL3NQ.
1/79 "Calibration-Spectrum Generator for the Microwave Bands up to 10GHz" by DK3UC.

2/79 "A Frequency Multiplier for Narrow Band 3cm Band Communications" by DK2VF and DJ1CR.
2/79 "A 3cm Primary Radiator for Parabolic Antennas" by DK2VF and DJ1CR.
2/79 "An SSB Transmitter for the 13cm Band" by VK5QR.
2/79 "Interdigital Converters as Transmit Mixers" by DK3UC.
2/79 "SSB-Transmit Mixers for the SHF Bands - Part 1: 13cm Band" by DC3QS.
2/79 "Technology and Frequency Plan for Repeaters in the 23cm Band" by DD0QT.

3/79 "A Frequency Doubler for the 13cm Band with 6W Output Power" by DF7QF.
3/79 "SSB Transmit Mixers for the SHF Bands - Part 2: The 9cm Band" by DC3QS.
3/79 "A Simple Radiator for 3cm Parabolic Dishes" by DC3QS.
3/79 "Optimum Spacings of Directional Antennas" by DL6WU.

4/79 "BIG WHEEL - An Omnidirectional Antenna for the 23cm Band" by DD0QT.
4/79 "A Tranceiver for the 10GHz Band" by DL6MH.

1/80 "Two-Stage Low-Noise Preamplifiers for the Amateur Bands from 24 to 12cm" by DJ6PI.
1/80 "Simple Design of Lambda/4 Stripline Circuits" by DC3CL.
1/80 "Receive Mixer for the 6cm Band" by DC3QS.
1/80 "SSB on the 10GHz Band - Information Regarding a Future Description in VHF Communications" by DC8UG and DB1PM.

2/80 "Determining the Sensitivity of Receive Systems with the Aid of Solar Noise" by DL6WU.
2/80 "Automatic Frequency Control and Suppression of Acoustic Feedback in Conjunction with 10GHz Tranceivers" by OE7WMI.

3/80 "SSB on the 10GHz Band - Part 1: Generation of the Local Oscillator Frequency" by DC8UG and DB1PM.
3/80 "Home-Made Parabolic Dishes for Microwave Applications" by DL6MH.
3/80 "Waveguide for the 24GHz Band" by DL3ER.

4/80 "Local Oscillator, Transmit Mixer, and Linear Amplifier for the 9cm Band" by DF5QZ.

1/81 "SSB on the 10GHz Band - Part 2: Waveguide Modules" by DC8UG and DB1PM.
1/81 "SSB on the 10GHz Band - Part 3: Intermediate Frequencies in the 2m or 70cm Band" by DC8UG and DB1PM.
1/81 "An ATV-Transmitter for the 24cm Band Constructed from Modules described in VHF Communications" by DJ4LB.
1/81 "Coaxial SHF-Connectors Constructed from Bicycle Tire Valves" by DL3ER.

2/81 "A Linear Amplifier for 1250MHz using the BFQ68" by DJ4LB.
2/81 "Constant-Amplitude PLL-SSB on the UHF and SHF Bands" by DF7QF.
2/81 "Chokes for Contactless Tuning of Waveguide Modules" by DL3ER.
2/81 "A New Method of Mounting and Feeding a Gunn Element using a BNC Connector" by DD0MQ

3/81 "A 1.3GHz Prescaler and Preamplifier for Frequency Counters" by DJ6PI.
3/81 "An Extremely Low-Noise 96MHz Crystal Oscillator for UHF/SHF Applications" by DK1AG.
3/81 "Coupled Microstriplines as Filter" by DC8EC.

4/81 "An Extremely Low-Noise 96MHz Crystal Oscillator for UHF/SHF

Applications - Final Part II" by DK1AG.
4/81 "A Linear 1W Power Amplifier for 2400MHz" by DC3QS.
4/81 "A Home-Made UHF/SHF Power Meter" by DF7QF.
4/81 "A Wavemeter for the Frequency Range 23.5 to 24.5GHz" by DL3ER.
4/81 "Line-of Sight Microwave Communications" by QE3HSC.

1/82 "A Gunn Oscillator for the 24GHz Band" by DC3QS.
1/82 "Some Pitfalls in Noise Figure Measurements" by G3YGF and G4FZZ.
1/82 "Stabilising the Operating Point of Preamplifiers and Linear Amplifiers" by DC8EC.

2/82 "A Receive Converter for the 6cm Band" by DD0QT.
2/82 "Experiments with a 10GHz Frequency Multiplier with Interdigital Filter Coupling" by DK3UC.
2/82 "A Straight-Through Mixer for 24GHz" by DL6MH.

3/82 "Extremely Long Yagi Antennas" by DL6WU.
3/82 "A Power Amplifier for 3456MHz Equipped with a YD1060 Tube" by DK1UV.
3/82 "Bias Voltage Circuit for Tubes of the 2C39/3CX100 Families" by LA8AK.
3/82 "A Gunn Oscillator, Detector, and Mixer Module for 24GHz" by DL3ER.

4/82 "A 6cm Transmitter for FM and SSB" by DF5QZ.
4/82 "Straight-Through Mixer for 24GHz" by DJ7FJ.
4/82 "A Spectrum Generator for the 24GHz Band" by DL3ER.

16.8.2

DUBUS

The following list contains items of interest to microwave enthusiasts which were printed in DUBUS during 1979, 1980, 1981 and 1982.

1/79 "Waveguide Mixer for 9cm" by DK1UV.
1/79 "Coaxial-to-waveguide adapter for 9cm" by DK1UV.
1/79 "18 el long-yagi for 13cm" by DC0BV.
1/79 "Dummy load and attenuator for 10GHz" by F1CV0.

2/79 "A 300mW Oscillator for the 1.2GHz Range" by DC9CS.
2/79 "A 6cm Waveguide Converter" by DB6NT.

3/79 "10W on 23cm with a Quad of 2N5944" by OZ3LQ.
3/79 "Experiences With Frequency Doublers" by DL7HG.
3/79 "Stability of the Gunn Oscillator SGX07" by M. Dontenwille.
3/79 "An Antenna Switch for the X-Band Region" by F1DMC.

4/79 "23cm Double Converter" by DL7HG.
4/79 "Back Feeded Parabolic Reflector Antenna" by F1DXI.
4/79 "10GHz Tests using Reflections from Obstacles" by F1AQS.

1/80 "GaAs FET Preamplifier for 13cm" by DJ8QL.
1/80 "A Waveguide Filter for 10GHz" by DB6NT.
1/80 "A 9cm Preamplifier With Two Stages" by DB6NT and PA0JGF.
1/80 "Varactor Tripler for 10GHz" by DK1UV.

2/80 "Multi-Band Radiator 1 - 12GHz" by DL7QY.
2/80 "Radiator Systems for Parabolic Reflectors" by DB6NT.
2/80 "A High Sensitive R.F. Meter" by HB9MIN.
2/80 "Troposcatter" by G3YGF.

3/80 "23cm Preamplifier with MGF1400" by DJ8QL.
3/80 "High Power Varactor Frequency-Multiplier" by DL7QY.

4/80 "Varactor Doubler for 2160MHz" by DC0DA.
4/80 "The 23cm 'Five-Minutes-Mixer'" by DD0QT.
4/80 "A Scanner for Gunnplexer Tranceivers" by LA8AK.

1/81 "A 9cm SSB Transverter" by DC0DA.
1/81 "Remote Tuning of a Gunn Cavity" by F1XL.
1/81 "Notes on the subject of 10GHz Transverters" by F6DLA.

2/81 "A Waveguide filter for 10GHz" by DC7BQ.

3/81 "Interdigital Bandpass Filter Calculation (100 - 8000MHz)" by DL7QY.

4/81 "6W Power Amplifier for 9cm" by DK1UV.
4/81 "GaAs FET Preamplifier for 13cm" by DC8UG.

1/82 "Varactor Quintupler for 6cm" by DC0DA.

2/82 "Omnidirectional X-Band Antenna with circular Polarisation" by DK2RV.
2/82 "13cm Transverter" by OE2PMJ.

3/82 "13cm Transverter" by DF5QZ.
3/82 "13cm Power Amplifier" by DC8UG.

4/82 "1.5GHz Divider" by DB9SB.
4/82 "Doppler Shift on EME" by OK1DAI.

16.8.3 Technical Items in the Newsletter 1980-82

We sometimes get requests for back issues from new subscribers. Ultimately a booklet containing all the technical items would be useful, but at present we can only supply copies of the originals. The following is a list of most technical items which appeared between 1980 and 1982 grouped into rough subject divisions. Back issues are available from G4KNZ, not RSGB HQ.

Wideband 10GHz:	
Making a Start on 10GHz	05/81
Mullard CL8630 X-Band Gunn Oscillator	06/82
Gunn Diode PSU	10/81
WB IF Preamplifier and Discriminator	06/82, 07/82
Horn Antennas from Oilcans	02/82
Waveguide Switch for 10GHz	10/81
Characteristics of WB FM	03/80
Doppler Tranciever for 10GHz	04/82
A Gunn Diode and Oscillator Test Rig	09/82
AFC System	02/81
Narrowband 10GHz:	
Alignment of G3JVL Mixers	04/80
G3JVL Image Recovery Mixer Mk II	01/81
5.7GHz JVL Transverter (including folded version)	09/81, 03/82
Multiplier & Mixer Diode Mount Improvements	01/82
Notes on the VHF Comms NB Tranceiver	05/82
Frequency Locking:	
Phase Locking of a Gunn	02/82
Problems using Droitwich Locked Frequency Standards	07/82
Local Oscillator Source:	
Oscillator Quality on 10GHz	05/80
Modifications to 384MHz Source	04/81
384MHz Source Temperature Compensation	07/81, 08/81
Upper Frequency Limit of Microwave Committee 384MHz Source	07/82
Microphone Preamp for 384MHz source	04/81
Low Voltage Drop Regulator	06/80
Test Gear:	
VHF/UHF Power Meter	05/80
Broadband Linear RF Power Meter	08/81
Audio Voltmeter	02/81
Notes on Switched Attenuators	10/82
High Power Directional Coupler	02/82
Semi-Rigid Directional Coupler	04/82
Notes on Using Directional Couplers	05/82
Calculating Bearings:	
Calculating Bearings	05/82
Notes on the National Grid and Conversion of Lat/Long to NGR	06/82, 07/82
NGR Program (Basic)	03/81, 06/81
Filters:	
Waveguide Band Pass Filters	08/82, 09/82
Waveguide Filters for 3.4GHz	08/82
1.3GHz Repeater Duplexing Filters	08/81
Waveguide Transitions	09/82
Intermediate Bands:	
Equipment Performance on the Intermediate Bands	08/82
3.4GHz High Level Mixer	04/81, 05/81
Waveguides for 3.4GHz	08/82
Antennas:	
Notes on Alford Slot Antennas	04/82
Balun for use with Semi-Rigid Coax	02/82
Use of Spherical Reflectors as Approximations to Paraboli	09/82
24GHz and up:	
Modifications to the GDO33	05/81
Improved 24GHz Diode Mount	03/82
Millimetric Bands and Waveguide sizes	10/81, 02/81

Note: Where two issues referred to under an item it is often important to look at both as the later one may be a correction to the first.

Radio Communication - 10GHz

The following is a list of items relating to 10GHz that have been published in Radio Comminication between 1970 and 1980, compiled by G4KGC.

10GHz Technical

Diode Mixer		May	71	335
Microwave Diodes		Aug	71	528ff
TWT Amplifiers		Nov	77	873
Klystron RF Source		Jan	74	26f
Klystron as RF Preamp		March	78	222
Directional Coupler (round hole for more 20dB coupling)		Nov	73	753
Putting together waveguide equipment (transceiver)		Oct	73	629f
Waveguide Dummy Loads/ Attenuators		Nov	72	741ff
Waveguide Variable Attenuators		Dec	72	820
Waveguide 16 to coaxial cable transition		Feb	75	131
Coax connectors for microwaves		Feb	77	129
Microwave Transistors		Jan	72	16f
Waveguides and Wavelengths		Dec	73	858
	and	Oct	96	757
Wideband Tranceiver		June	78	492f
MA Gunnplexer, Equipment Review		Nov	78	930f
MA Gunn Transceiver		March	77	201
Gunn Transceiver, improved		Oct	77	791
Simplified Gunn Oscillators		Feb	76	123
Tuneable Gunn Oscillators		Sept	76	667ff
Gunn Oscillators		Aug	73	561
Gunn Diodes		April	71	271
Gunn Diode PSU's		Dec	76	919
Gunn Diode PSU and Modulator		March	72	158
30MHz IF amplifier for microwave receivers		July	72	430
GM3OXX Portable Transceiver		June	75	450ff
30MHz Wideband FM Discriminatpr with afc output		March	73	193
A Simple Polarplexer		Nov	70	752f
Polarization and the Polarplexer		March	71	188
G3JVL image recovery mixer/transmitter		Jan	79	41
Corrections to JVL		April	79	342
Further info to JVL		April	80	372
SSB		Dec	78	1037
Cavity Locking of Gunn Oscillators		Jan	78	52f
Varactor Multiplier		March	76	202
Low Power Varactor Multiplier		July	75	549
Crystal Controlled solid state source		Aug	75	614f
Injection Locking		June	73	398
Injection Locking for reflex Klystron Oscillator		Nov	74	754ff
Filters		Aug	73	560f
G3JVL bandpass filter		Oct	77	791
Tone modulation of local oscillator		Feb	72	97
IF Frequencies		March	71	188f
	and	July	73	478
Polarization		Jan	71	19
Microwave components		March	78	228
Broadband and narrowband		Feb	72	81f
GHz, MHz, cms and Bands		Jan	71	19

Antennas for 10GHz

Optimum Horn Design		Feb	72	81
Correction to Optimum Horn Design		April	72	229
Designing Paraboloids		April	71	214ff
Saucepan lid dish		FEb	75	125
Dustbin lid antenna		Oct	74	691
Dustbin lid dish and Feed		March	76	194ff
Feedhorns for dishes		April	72	229ff
	and	March	75	187ff
Waveguide Feed for shallow dishes		Oct	76	757
A Tripod for Dishes		March	80	272
A Mast system for dishes		May	73	310ff
Flyswatter antenna		Aug	77	596ff
G3JVL Loop Yagi		Jan	78	53

168.4 page 2

10GHz Testing and Setting Up				
G4COM receiver alignment aid		Jan	76	
	and	May	79	424
	and	Sept	79	832
Debate on COM		Feb	80	155
Modifications to COM		March	80	270f
Optimising microwave receivers		Oct	79	934
Checking microwave receiver performance		July	77	541
System checking with sun noise		July	79	636
Moon noise		April	78	333
Frequency checking		Oct	71	683
Crystal Controlled frequency marker		May	76	352f
Droitwich standard ideas		Feb	79	134f
Self Calibrating Wavemeter		Feb	73	106
High Q Wavemeter		April	73	261
Rough Measurement of Power		Sept	73	618
Schottky diode power-current relationship		Nov	73	753
Equipment Performance				
Calculating Equipment Perfomance		March	78	222ff
Equipment Performance over Line-of-sight paths		July	72	442
	and	May	78	413
Equipment Performance over Non-line-of-sight paths		July	78	610f
Power required for Troposcatter		May	76	363
10GHz Operating				
Take to the hills - operating /P		July	72	436ff
Sun for aligning antennas		June	79	536f
Pointing high gain antennas		Sept	74	592f
Microwave path checking		Feb	78	122ff
Operating standards for microwave bands		June	71	391
Setting up microwave contacts		Jan	72	15
Microwave calling frequency		July	73	560
	and	May	80	493
10GHz operating/talkback polarisation		Feb	78	135
Moving wideband to narrowband section		May	73	325
	and	July	73	473
Listening to GB3LEX mobile		May	80	493
SSB mobile		Aug	79	744
TV		Aug	72	529
Moonbounce attempt		July	75	549
Radiation hazards		July	71	467
GB3IOW		May	75	389
	and	Sept	75	686
GB3LBH		July	76	525
PA0HSM		July	77	540
UK Microwave beacons		Jan	78	53
10GHz Progagtion				
Meterological aspects of anomalous propagation		Feb	78	126ff
Propagation losses over non-optical paths		July	78	610f
Superrefraction		Sept	73	618f
	and	Nov	74	772
Superrefraction, conditions for ...		Sept	74	592
Overland superrefraction		Nov	75	852
Superrefraction vs time of day		Oct	75	788
Overseas enhancements during bad weather		Sept	78	782
Troposcatter in weatherfronts, inversion, haze		Nov	78	950f
SSB, rapid flutter and deep fading		Dec	78	1037
Weatherfront ducting		Feb	79	140
Wind, weatherfronts		May	79	435
Enhancements in fog, thunderstorm and weatherfronts		Sept	79	850
Weatherfront, K-factor, superrefraction		Nov	79	1044
Aeroplane Scatter		Nov	79	1044
Tropo openings, "lifts"		March	80	270
	and	April	80	382
Unusual signls with extremely rapid 'chirpy' fading		May	80	492